Electronic technology

principal contents

Telemetry Signals from Sputnik III

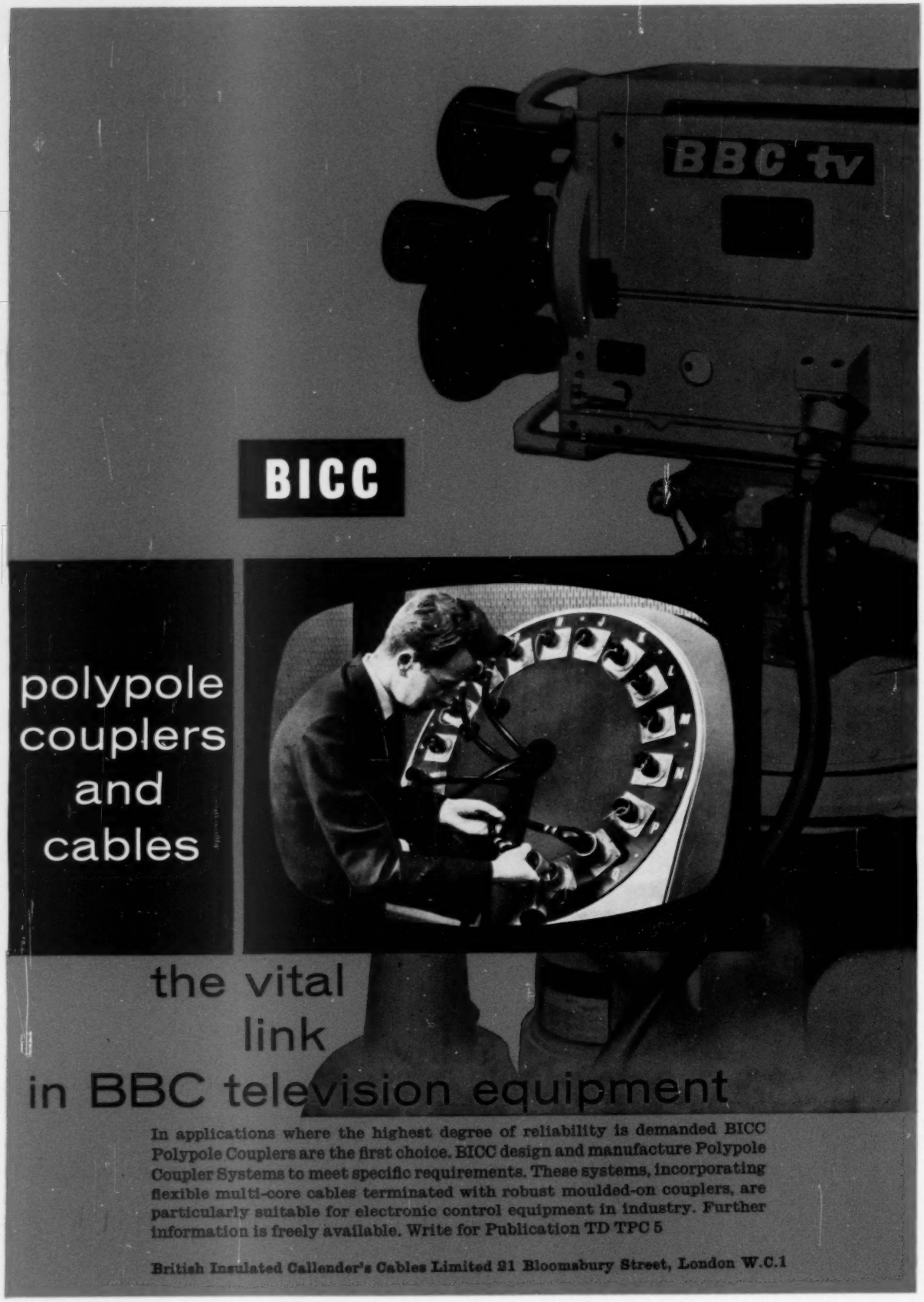
Transistor Frequency Response

Barkhausen Noise in Transformer Cores

Theory of Waveguides and Cavities

FIVE SHILLINGS

MARCH 1961 VOLUME 38 NUMBER 3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polypole couplers and cables
the vital link in BBC television equipment
In applications where the highest degree of reliability is demanded BICC Polypole Couplers are the first choice. BICC design and manufacture Polypole Coupler Systems to meet specific requirements. These systems, incorporating flexible multi-core cables terminated with robust moulded-on couplers, are particularly suitable for electronic control equipment in industry. Further information is freely available. Write for Publication TD TPC 5
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Noise power excluding image frequency contribution	15·25db
Operating Current	35mA
Overall Length	$6\frac{1}{32}$"
Base Diameter	0·64"
Discharge Tube Diameter	0·185"

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Ferranti

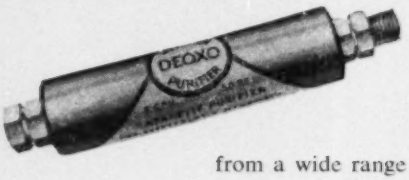
DS/T73

OXYGEN REMOVAL and MEASUREMENT

For the removal of oxygen from other gases

DEOXO CATALYTIC GAS PURIFICATION PROCESS

The model 'D' Deoxo Purifier, which operates at room temperature, provides for the removal of oxygen or hydrogen from other gases such as argon, helium nitrogen, etc., with a remaining impurity of less than 1 ppm. These Purifiers are available in a number of standard sizes for gas flows from 5 scfh to 10,000 scfh.



Additional units are available for the removal of other impurities from a wide range of industrial gases.

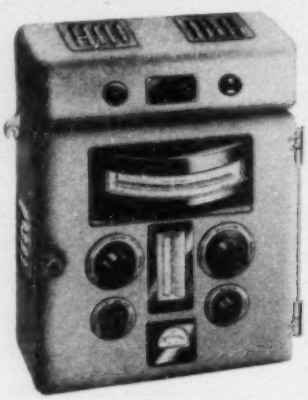
For pure dry hydrogen

DEOXO PURIDRYER

The Deoxo Puridryer will remove oxygen from hydrogen to leave less than 1 ppm of the impurity remaining, and subsequently dry the purified gas to a dew point as low as minus 100° F, at flow rates of up to 100 scfh.

Provision for regenerating the drying agent is incorporated in the unit.

For continuous measurement of oxygen



HERSCH OXYGEN METER

The Hersch Oxygen Meter has been designed for the measurement of trace oxygen in other gases in the range 0-10 and 0-100 ppm. The high sensitivity, accuracy and speed of response enables this instrument to be used both for laboratory investigation and production quality control.

★ *Illustrated literature available on request.*

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52 HIGH HOLBORN · LONDON · W C.1 · *Telephone: CHAncery 8711 & 6506*

We can get you out of an Audio-Valve problem.

Here is a brilliant newcomer to the celebrated KT series of beam tetrodes—the KT 77. We have been getting people out of valve problems for over 40 years and now we've put all that experience into this valve, which joins our KT 55, 66 and 88 range.

	P_{out}† (W)	Distortion† (%)	V supply (V)	DIMENSIONS Seated height (mm)	DIMENSIONS Max diameter (mm)
KT 55	25	1.5	*	128	53
KT 66	50	3	525	120	53
KT 77	70	1	660	98	33
KT 88	100	1.5	500	110	53

† Class AB_1, ultra-linear connection
* 190—260 V, AC or DC mains

A comprehensive range of audio valves is available, to give outputs up to at least 50 kW.

G.E.C.

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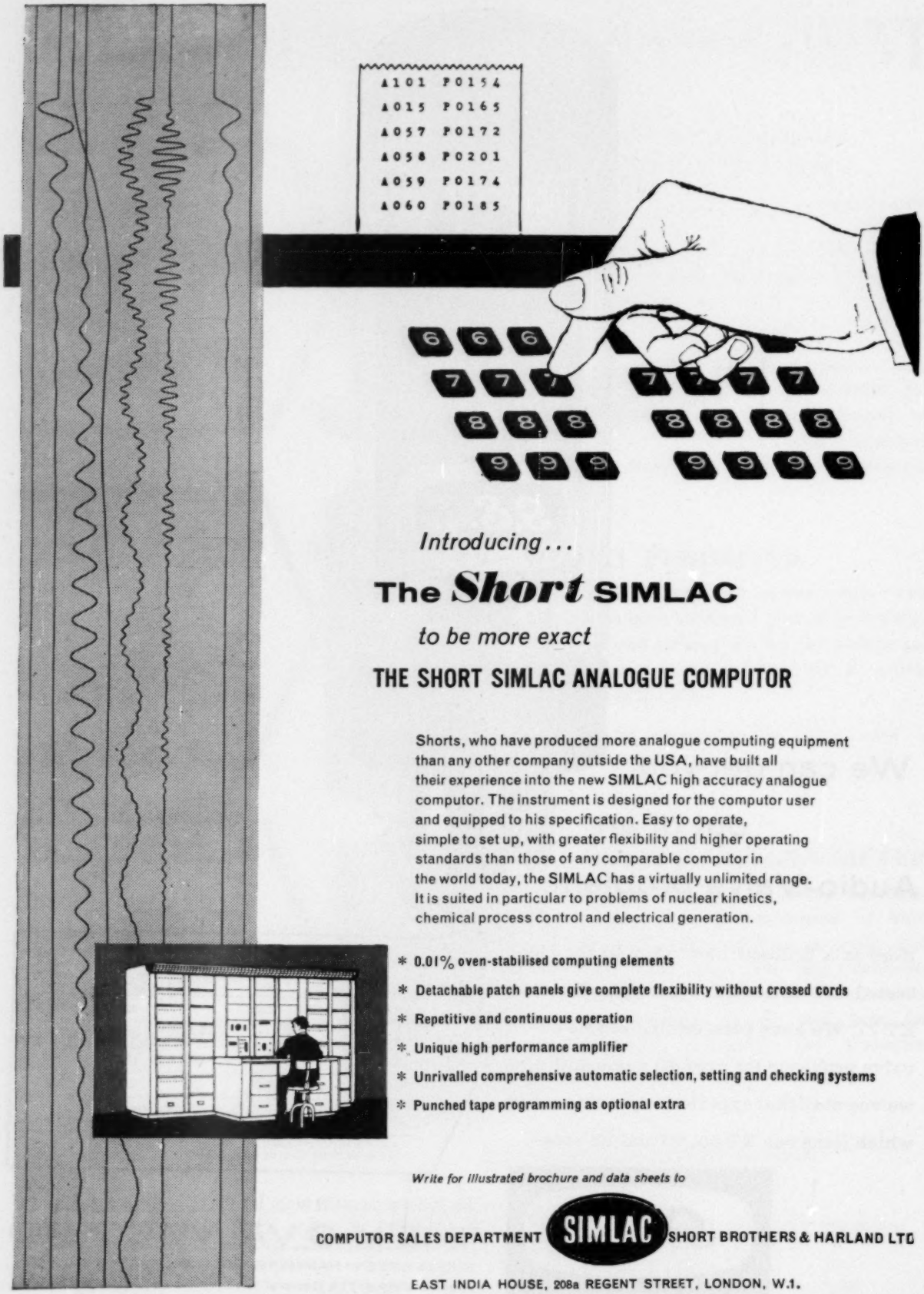

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THE SHORT SIMLAC ANALOGUE COMPUTOR

Shorts, who have produced more analogue computing equipment than any other company outside the USA, have built all their experience into the new SIMLAC high accuracy analogue computor. The instrument is designed for the computor user and equipped to his specification. Easy to operate, simple to set up, with greater flexibility and higher operating standards than those of any comparable computor in the world today, the SIMLAC has a virtually unlimited range. It is suited in particular to problems of nuclear kinetics, chemical process control and electrical generation.

* 0.01% oven-stabilised computing elements
* Detachable patch panels give complete flexibility without crossed cords
* Repetitive and continuous operation
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* Unrivalled comprehensive automatic selection, setting and checking systems
* Punched tape programming as optional extra

Write for illustrated brochure and data sheets to

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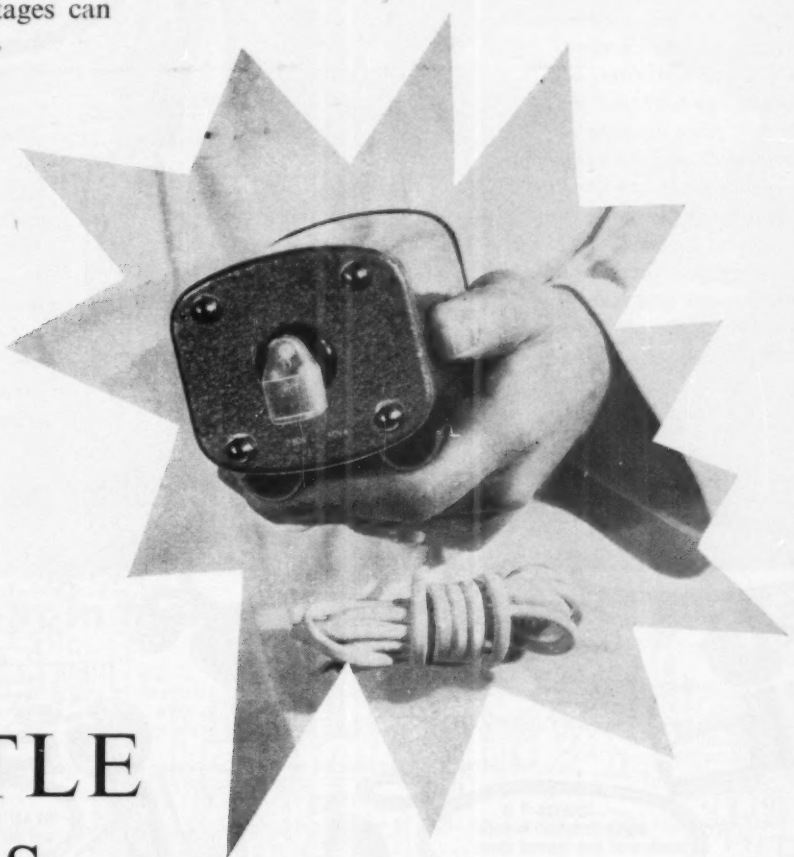

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TRANSFORMERS

22 standard double wound ratings and 9 auto wound are now listed, with dimensions, weights and prices in Gardners Industrial and 'SM' (Private users) Catalogues. Existing registered holders of the Industrial Catalogue have been supplied with Amendment Sheets covering the new Castle Series.

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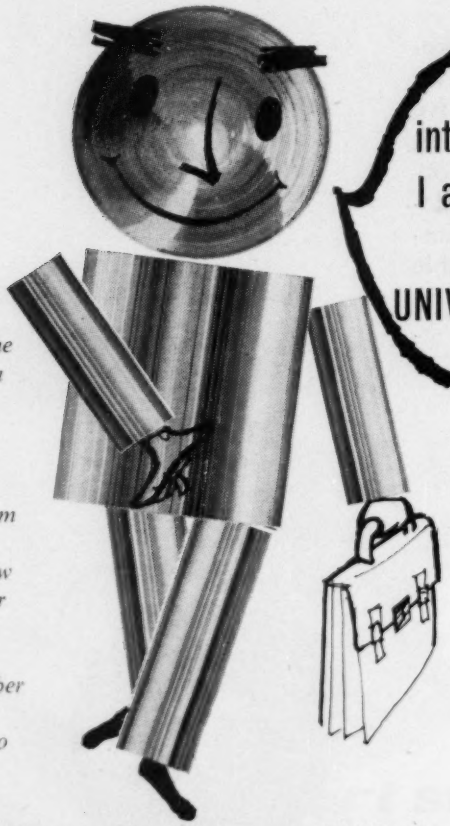
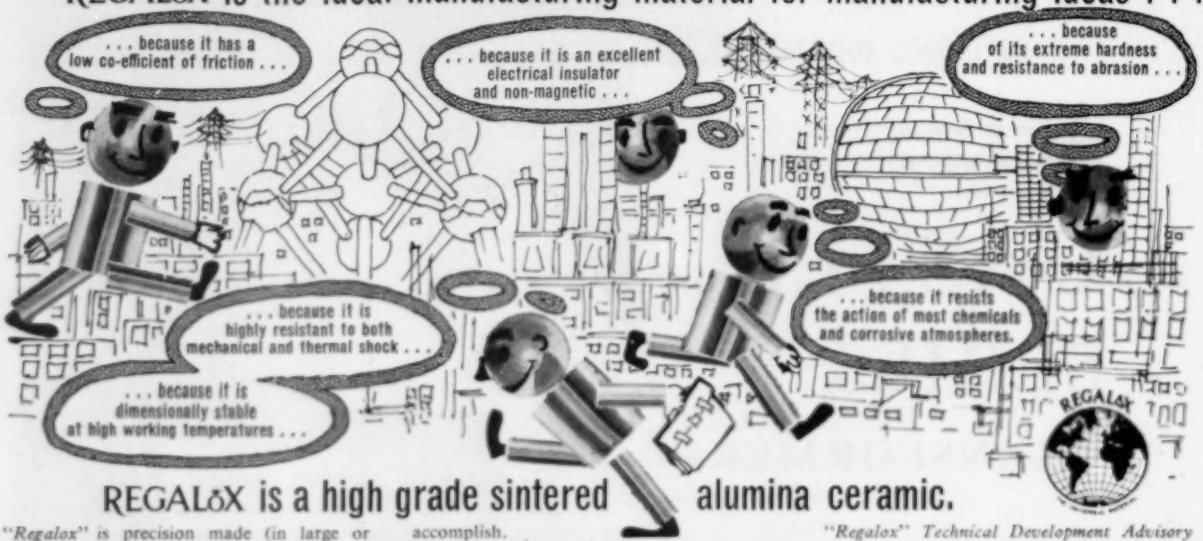


let me introduce myself... I am REGALOX the UNIVERSAL MATERIAL

I was born in South Wales during the second World War and come from a celebrated family—my ancestors being pure Royal Worcester. In a way I suppose you might call me a son of the soil, since I originate from aluminium oxide, but right from my early youth I was attracted to modern industry and with this in view I was brought up the hard way. Over the years research workers have tried to find a weak spot in my make-up and wear me out in a number of ways. They've also boiled me in corrosive chemicals—thrown me into fiery furnaces—tried to make me drink (to no avail since my absorption rate is nil) and subjected me to high voltage electrical currents without breaking me down.

They still keep at me, of course, as new ideas in industry make more stringent demands on materials—particularly in my case—but now they treat me with new respect and freely admit that in me ***you*** *have a hard-working, reliable servant for all forms of industry.*

REGALOX is the ideal manufacturing material for manufacturing ideas . . .


. . . because it has a low co-efficient of friction . . .
. . . because it is an excellent electrical insulator and non-magnetic . . .
. . . because of its extreme hardness and resistance to abrasion . . .
. . . because it is highly resistant to both mechanical and thermal shock . . .
. . . because it is dimensionally stable at high working temperatures . . .
. . . because it resists the action of most chemicals and corrosive atmospheres.
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REGALOX is a high grade sintered alumina ceramic.

"Regalox" is precision made (in large or small quantities designed to meet your specific requirements).

"Regalox" is used for a wide variety of Electrical, Mechanical, Chemical and Engineering applications and because of it's unique characteristics successfully and efficiently performs many jobs no other material can accomplish.

"Regalox" will serve you better than any other material if you have a need for a non-metallic material that must be hard, strong and smooth, requiring exceptional resistance to extreme temperatures, corrosion and abrasion, yet capable of being produced economically to close tolerances.

"Regalox" Technical Development Advisory Service is freely available for advice on how Regalox can be applied to your product or production and maintenance problems. A rapid service is provided from the range of standard stocks (normal as fired-finish or diamond-ground or polished) and quotations are given, without obligation.

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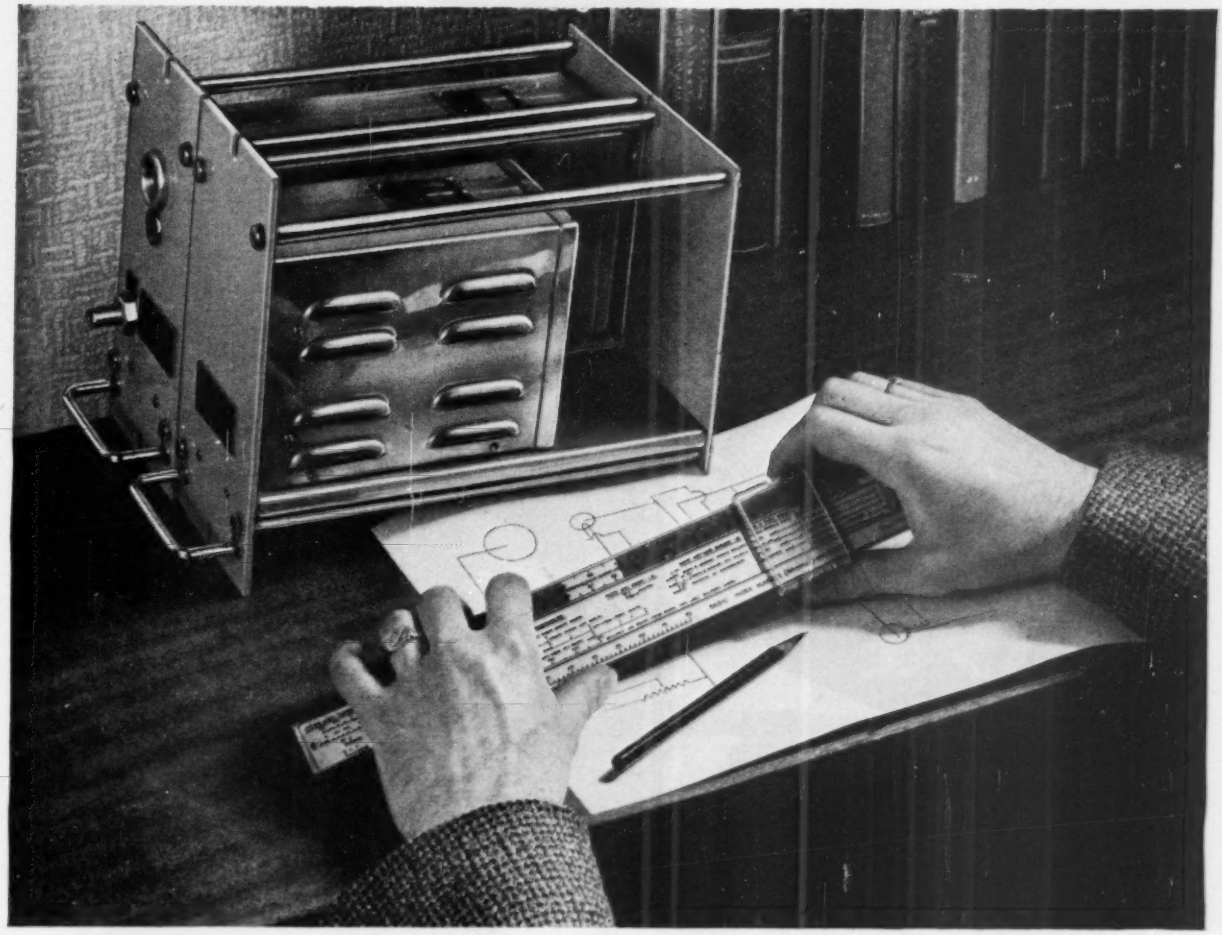
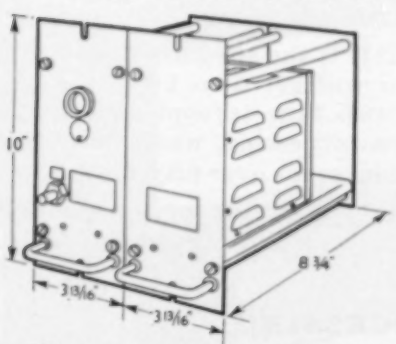


a new concept in high-stability frequency sources

The latest range of compact, high-stability frequency sources by STC represents considerable progress in size, weight and price reduction without sacrifice of ageing characteristics, short-term stability or temperature dependence.

Illustrated is the 90 LQU 802B comprising type 57 LQU 423 Quartz Crystal Unit and Temperature Control Unit Type 1LQU 101B :—



STABILITY	± 4 in 10^{10}
AGEING	1 in 10^{9} per day
DIMENSIONS	Two Units 10 in. x 8¼ in. x 3½ in.

Full details of all units in the range available on request from —

STC COMPONENTS GROUP

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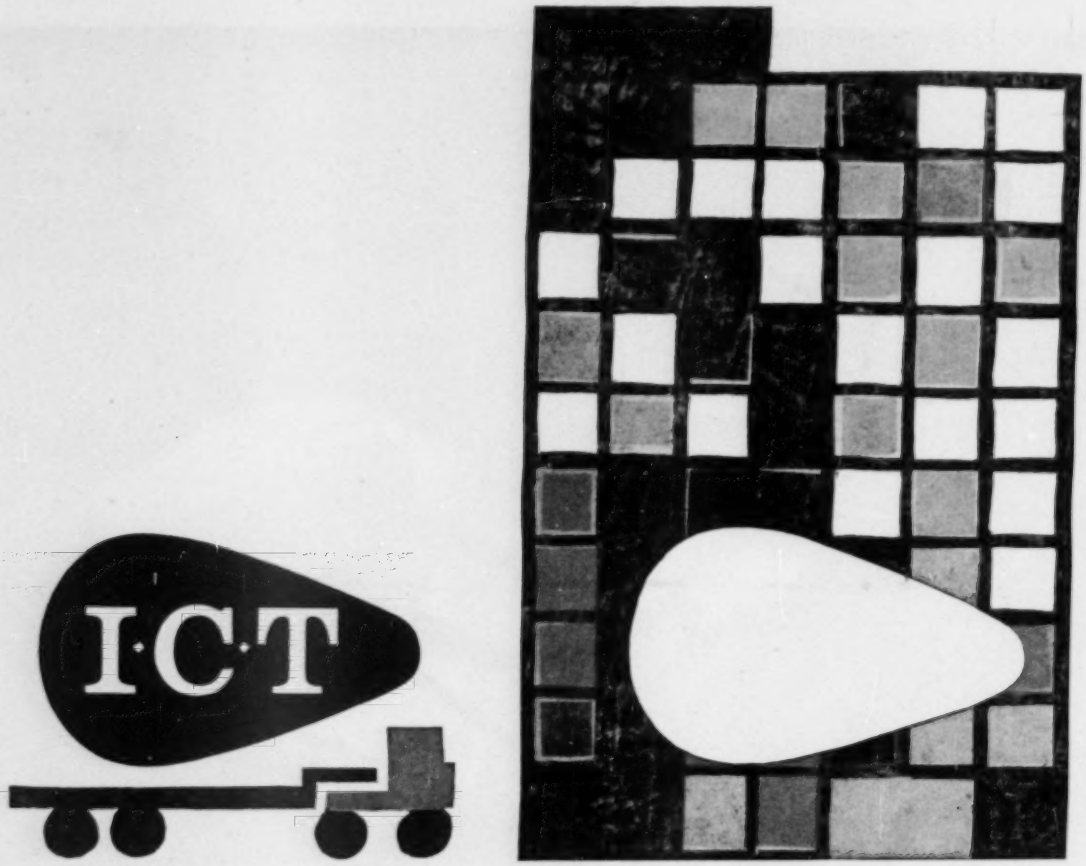

There's an I·C·T computer to fit your business

Shopping for a computer? Then you've a challenging task ahead. For this major investment must pay you major dividends.

Choosing and using a computer effectively and profitably requires imagination and great preparation. This means thinking in terms of the service you'll get as well as the machines you will use. Doing your shopping, in fact, with an organisation having the 'know-how', the resources and the experienced staff to examine your problems creatively, and to propose the data processing system that *fits into your business*.

This is the specialised service that you will get from I·C·T—the largest British firm devoted exclusively to data processing, which has been its business for over fifty years.

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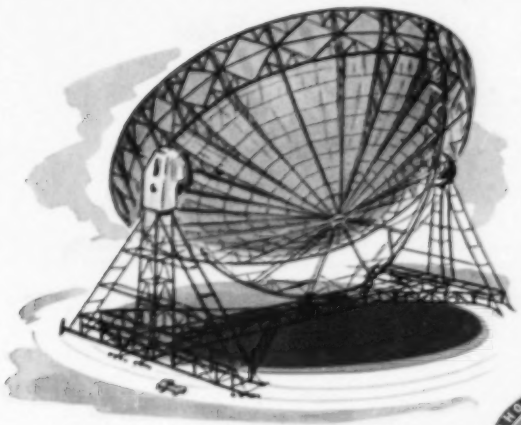
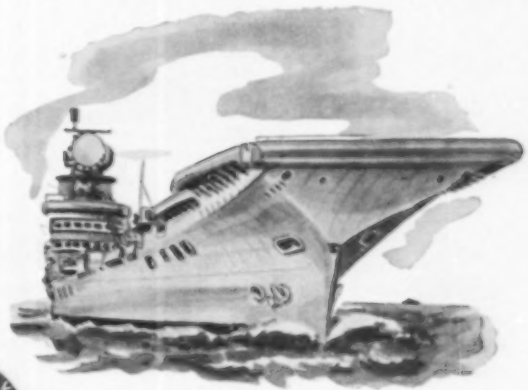
For Service & Professional Communications





Superbly engineered and of advanced design, the two models offered possess excellent electrical characteristics and are robustly constructed for service in any climate. The "770 R" has continuous AM/FM coverage from 19 Mc/s to 165 Mc/s; the "770 U" from 150 Mc/s to 500 Mc/s. Both incorporate six-position turret tuning assemblies of unique design and giving high reliability. Illustrated literature with full information and performance curves sent on request.

MANUFACTURERS

STRATTON & CO. LTD · BIRMINGHAM 31

SILVER STAR square section CAPACITORS

for transistorised circuits

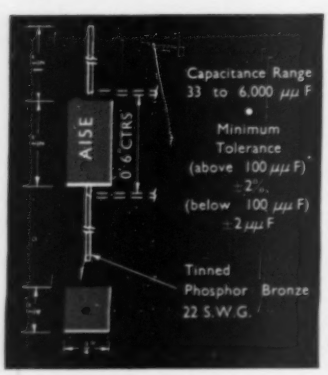
Something new in silvered mica capacitors — new in the Silver Star range!

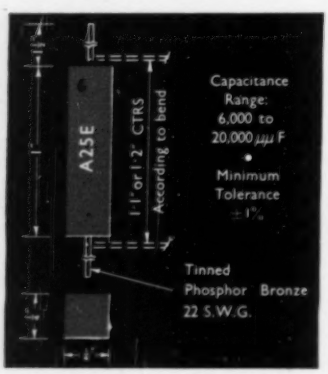
Designed specially for transistor circuitry, these miniature square-section capacitors are particularly useful when the component density is high.

And they are ideal for printed circuits, for they can be cemented to the circuit board — a big advantage when acceleration forces are high.

Two sizes (A15E and A25E) together cover the capacitance range from 33 to 20,000 μμF. For values above 6,000 μμF the capacitance tolerance can be as close as ±1 per cent.

- **Square section**
- **Miniature size**
- **Ideal fc rinted circuitry**
- **50 volt rating**
- **Fired construction for maximum long-term stability**
- **Synthetic resin encapsulated**
- **Ideal for "high g" applications**
- **Silver Star © quality**





Full technical information available on request.

One of the specialised products of

Johnson Matthey

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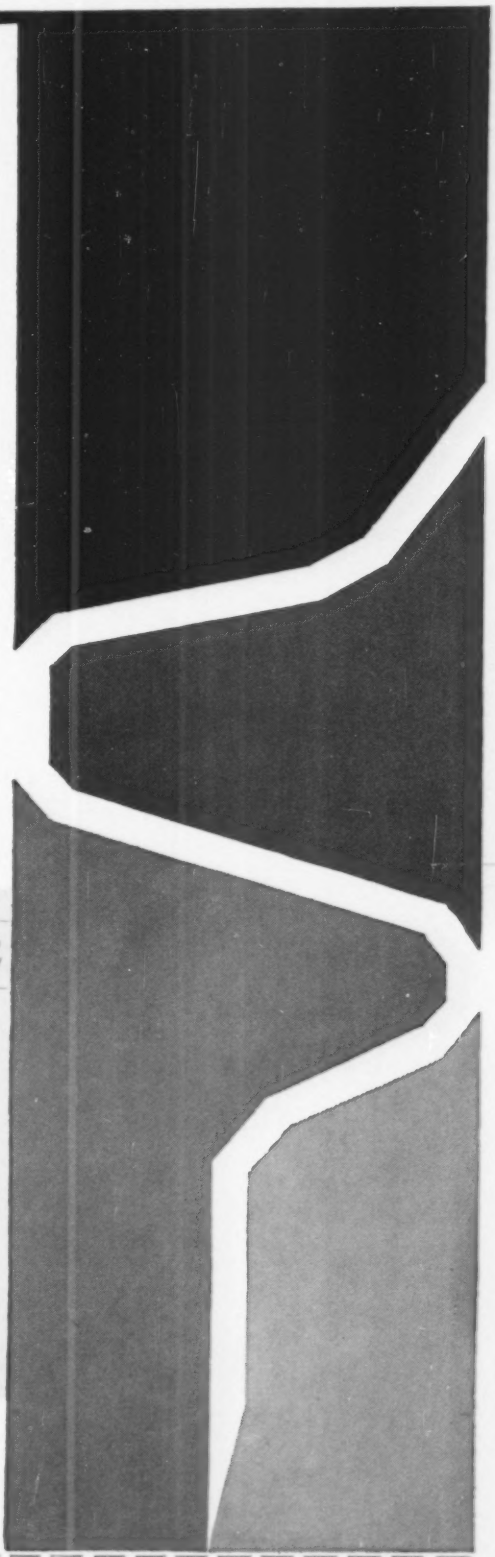
specify H point contact germanium diodes



Hughes HG10 series are point contact germanium diodes with high peak inverse voltage, high back resistance and good forward characteristics. They have excellent recovery characteristics and evidence no forward current saturation within their power limitations.

Made in Glenrothes, Scotland, the Hughes range of silicon and germanium diodes are subminiature devices with extremely stable electrical and mechanical characteristics. These diodes are specially designed and constructed to meet the most exacting requirements of military or commercial applications. They are double wire ended and fusion-sealed in a subminiature one-piece glass envelope to ensure complete isolation of the active elements from damage or contamination. The small size, combined with rigidity of construction and small mass of the elements, enable them to withstand successfully physical shock and vibration.

Type No.	Continuous Ratings 25°C		Characteristics @ 25°C	
	Peak Inverse Volts	Max. Forward Current	Min. Forward Current at 1V	Max. Reverse Current at 50V
HG1001	100	45 mA	20 mA	50 μa
HG1002	100	45 mA	20 mA	100 μa
HG1003	100	45 mA	10 mA	50 μa
HG1004	100	45 mA	10 mA	100 μa
HG1005	100	45 mA	5 mA	50 μa
HG1006	100	45 mA	5 mA	100 μa
HG1007	75	45 mA	20 mA	50 μa
HG1008	75	45 mA	20 mA	100 μa
HG1009	75	45 mA	10 mA	50 μa
HG1010	75	45 mA	10 mA	100 μa
HG1011	75	45 mA	5 mA	50 μa
HG1012	75	45 mA	5 mA	100 μa

Actual Size

Qualified engineers in our Research and Development laboratories at Glenrothes are available to help with your application problems.

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magnétophone professionnel haute fidélité

high fidelity tape recorder F 100 for profes-sional use

F 100

2 tape speeds
(3 3/4″ and 7 1/2″ or
7 1/2″ and 15″ per second)
Can be used with or
without microphone preamplifier,
mixer and monitoring amplifier.
Type F 100 M : multi-track (2 tracks : stereo)
Overall response curves F 100 and F 100 M :
15″ per second : 40 to 15,000 c/s ± 2 dB
7 1/2″ per second : 40 to 15,000 c/s + 2 −4 dB

LE Belin

2 vitesses (9,5 et 19 ou 19 et 38 cm/s)
peut être utilisé avec ou sans pré-ampli micro,
ampli de mélange et de contrôle.
Type F 100 M : multi-pistes (2 pistes : stéréo)
Courbes de réponse globales F 100 et F 100 M :
38 cm/s : 40 à 15 000 Hz ± 2 dB
19 cm/s : 40 à 15 000 Hz + 2 − 4 dB

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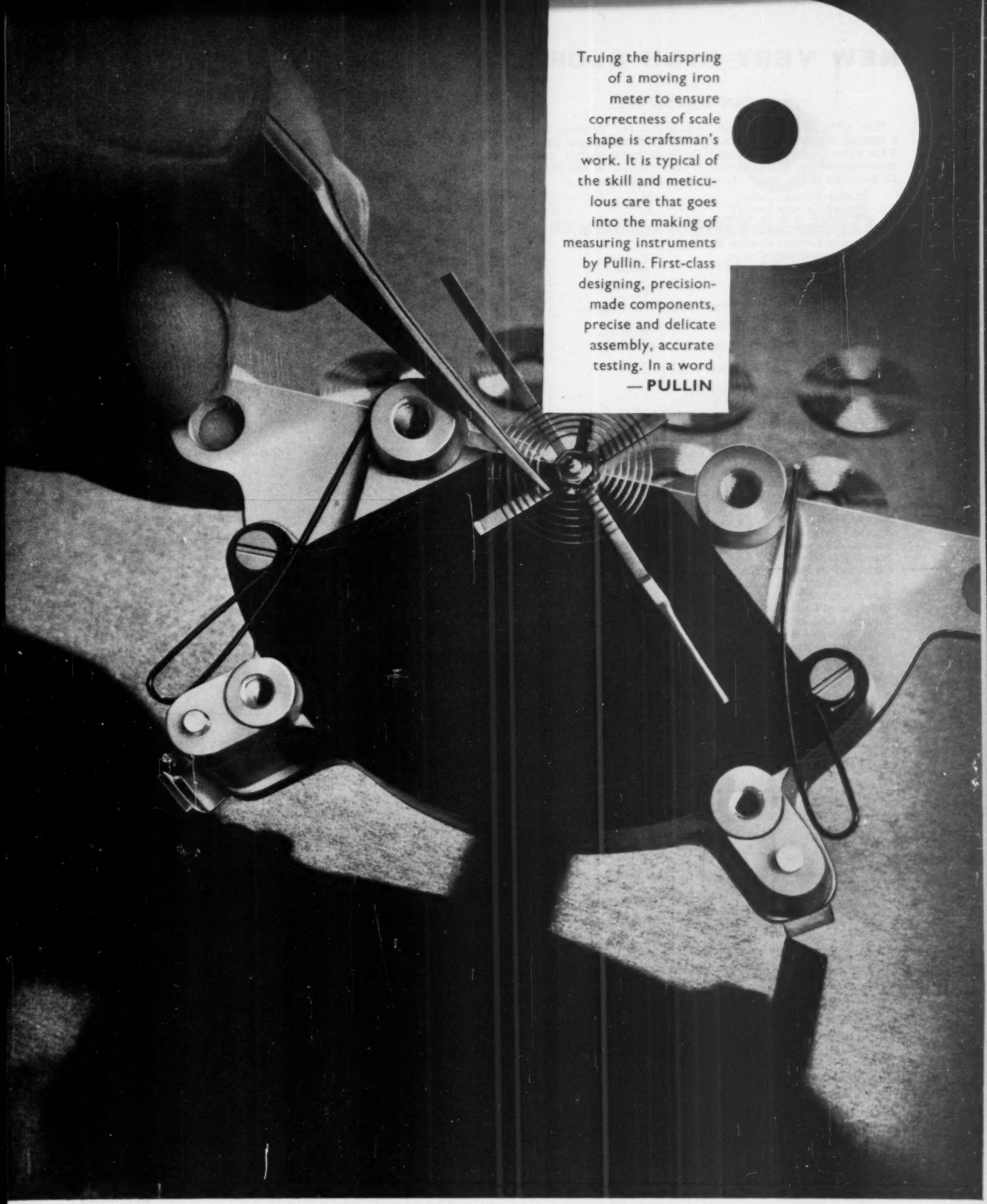
296, Avenue Napoléon-Bonaparte - RUEIL-MALMAISON (Seine-et-Oise France) - Tél. 967-15-54



Truing the hairspring of a moving iron meter to ensure correctness of scale shape is craftsman's work. It is typical of the skill and meticulous care that goes into the making of measuring instruments by Pullin. First-class designing, precision-made components, precise and delicate assembly, accurate testing. In a word — **PULLIN**

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NEW VERY HIGH SLOPE VARIABLE-MU HF PENTODE

EDISWAN MAZDA 6F25

The 6F25 is a frame grid variable-mu HF pentode having a mutual conductance of 12.5mA/V at 11.5mA anode current. This valve is capable of handling large input signals without cross-modulation and is therefore very suitable as the common IF amplifier in a.c. or a.c./d.c. television receivers.

Heater Voltage (volts)	V_h	6.3
Heater Current (amps)	I_h	0.3

TENTATIVE RATINGS AND DATA

Maximum Design Centre Ratings

Anode Dissipation (watts)	$p_{a\ (max)}$	2.5
Screen Dissipation (watts)	$p_{g2\ (max)}$	0.5
Anode Voltage (volts)	$V_{a\ (max)}$	250
Screen Voltage (volts)	$V_{g2\ (max)}$	250
Heater to Cathode Voltage (volts rms)	$V_{h-k\ (max)\ rms}$	150*
Resistance Control Grid to Cathode (megohms)	$R_{g-k\ (max)}$	1

* From cathode to higher potential heater pin.

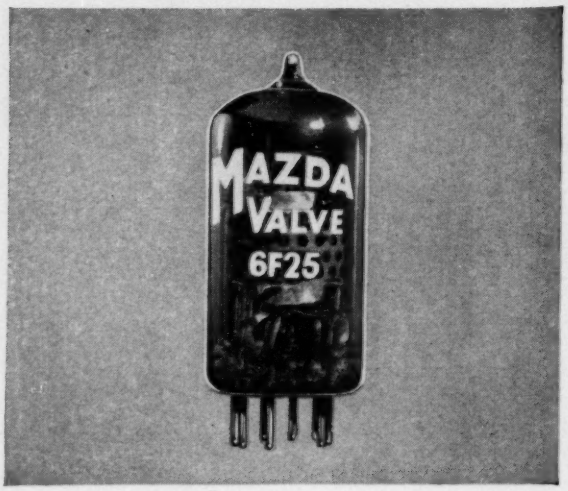
Inter-Electrode Capacitances† (pF)

Input Capacitance	C_{in}	8.5
Output Capacitance	C_{out}	2.7
Grid 1 to Anode	C_{g1-a}	0.006
Grid 1 to Grid 3	C_{g1-g3}	0.1
Grid 1 to Grid 2	C_{g1-g2}	1.8
Grid 1 to Cathode	C_{g1-k}	6.0
Grid 2 to Anode	C_{g2-a}	0.19
Grid 3 to Anode	C_{g3-a}	0.45

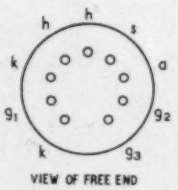
† Measure in fully shielded socket, without can.

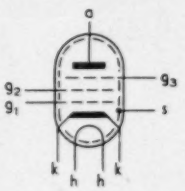
TYPICAL OPERATION

Supply Voltage (volts)	V_b	200
Anode Voltage (volts)	V_a	170
Screen Voltage-Initial (volts)	V_{g2}	90
Screen Resistor (kΩ)	R_{g2}	39
Cathode Bias Resistor (ohms)	R_k	100
Anode Current (mA)	I_a	11.5
Screen Current (mA)	I_{g2}	2.8
Inner Amplification Factor (g_1 to g_2)	μ_{g1-g2}	35
Mutual Conductance (mA/V)	g_m	12.2
Grid Voltage for Mutual Conductance = 1.25 mA/V (volts)	—	−5.5



Base: B9A (Noval) **Mounting Position:** Unrestricted

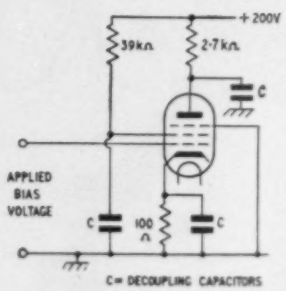




Tentative Characteristic Curves of Ediswan Mazda Valve Type 6F25

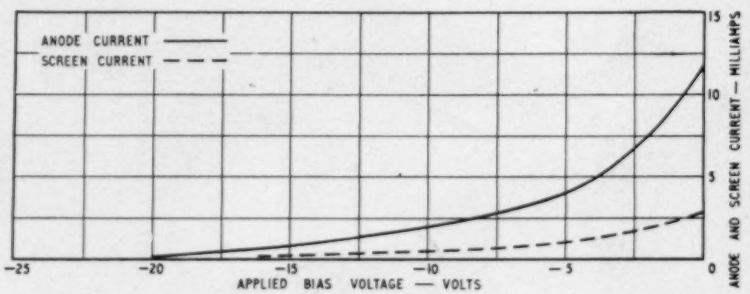
RISING SCREEN CHARACTERISTICS

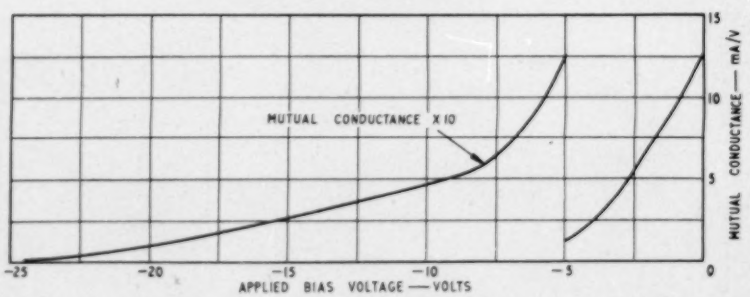
TEST CIRCUIT



Maximum Dimensions (mm)

Overall Length	56
Seated Height	49
Diameter	22.2





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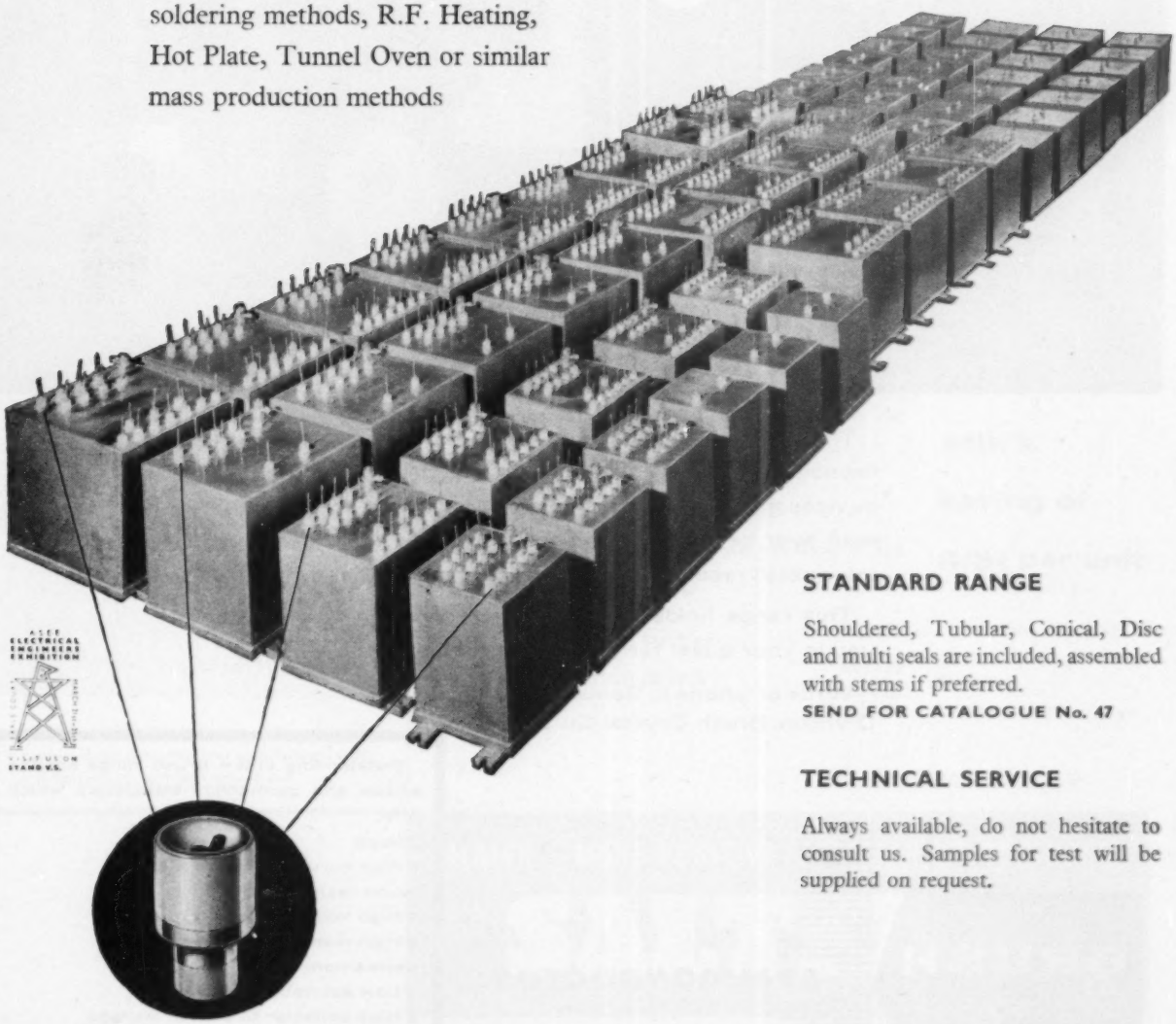

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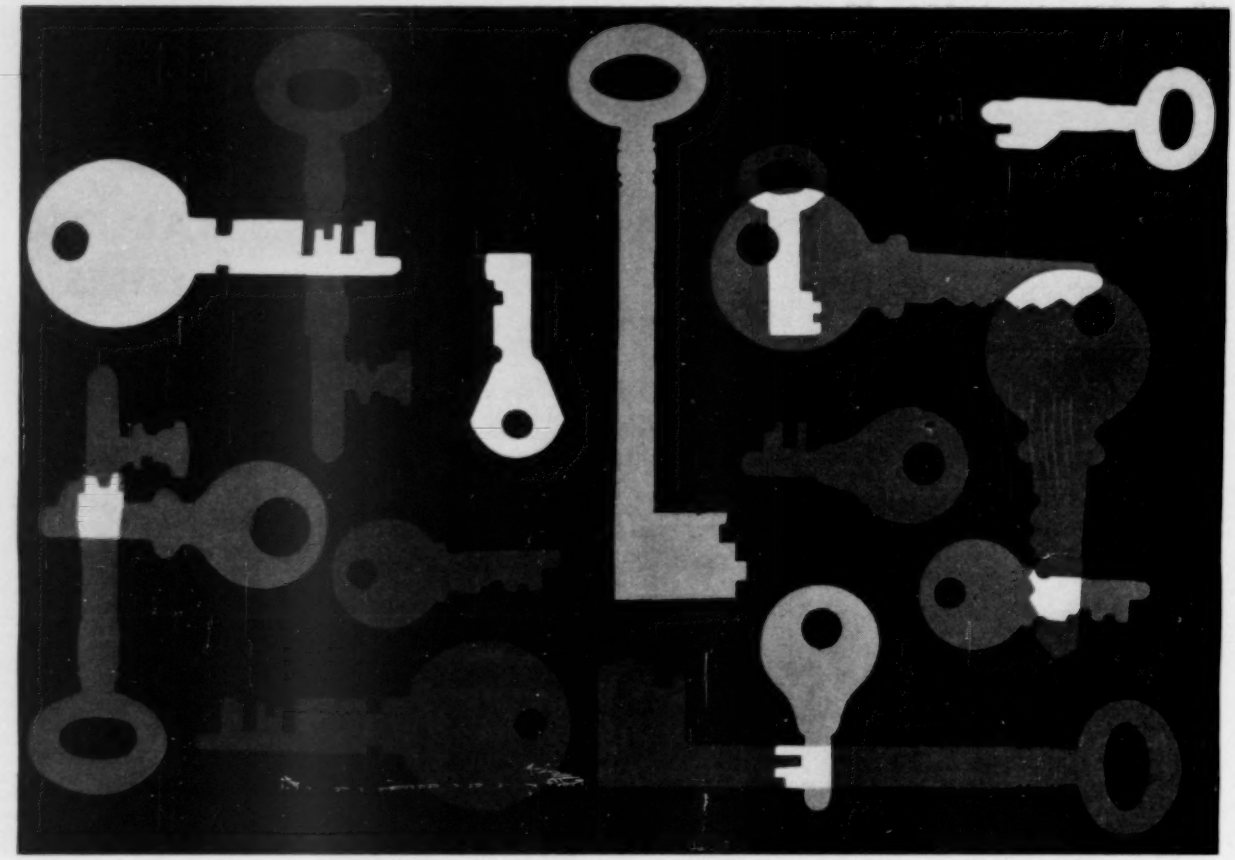

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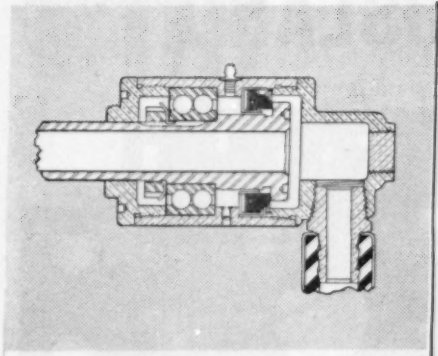

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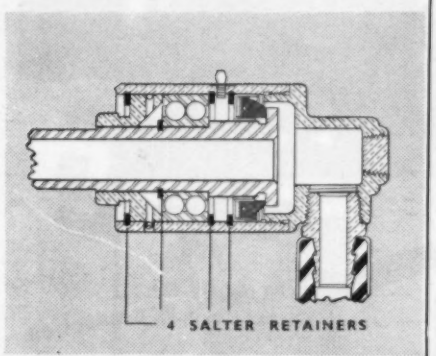

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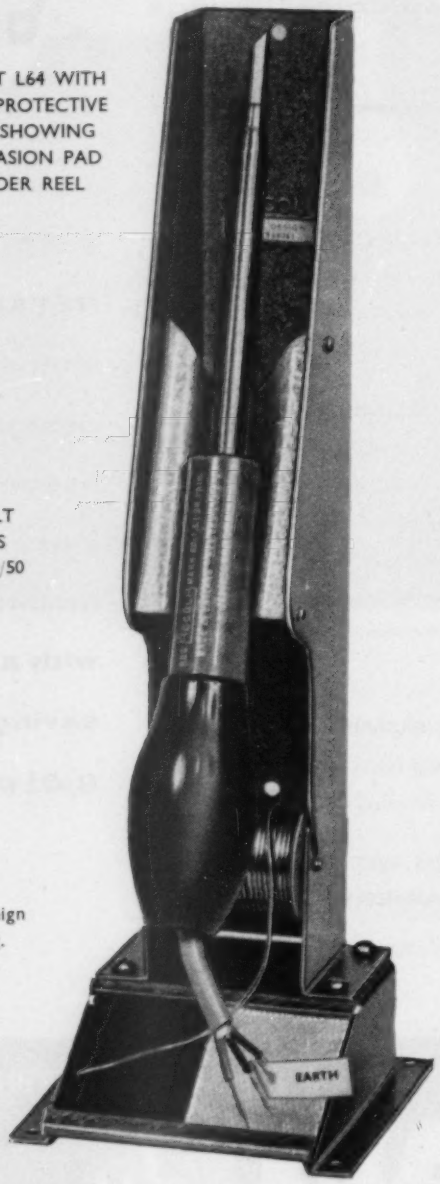

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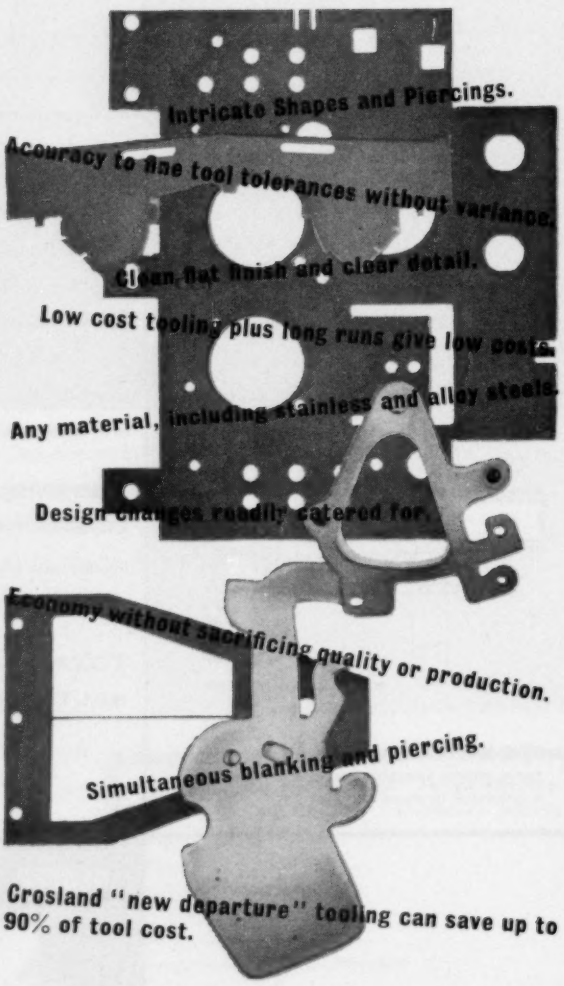
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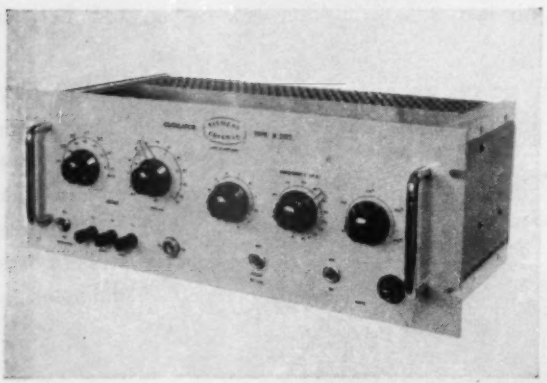

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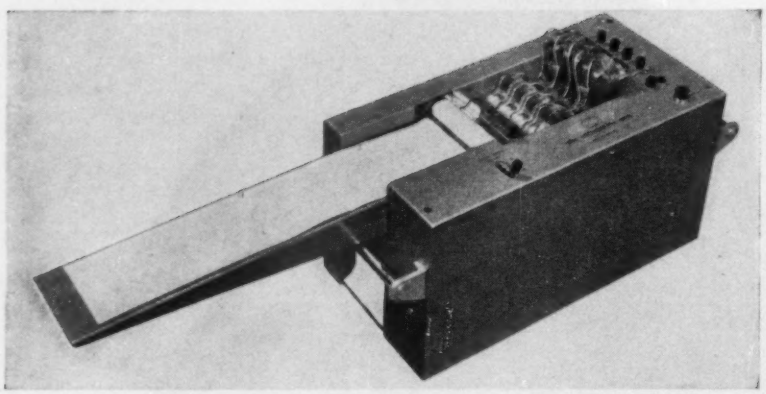

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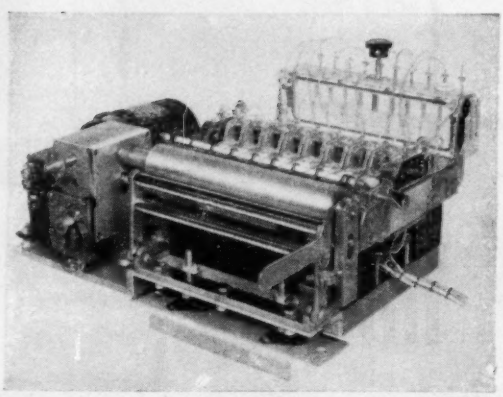

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CONTENTS

VOLUME 38 NUMBER 3 MARCH 1961

Transistor Mutual Conductance	**75**	Editorial
Telemetry Signals from Sputnik III	**76**	*by R. E. Henderson, B.Sc.*
Transistor Frequency Response	**80**	*by J. R. James and D. J. Bradley*
The Fringe of the Field	**83**	*by Quantum*
A New Look at the Transistor Symbol	**87**	*by R. van Winkle*
Barkhausen Noise in Transformer Cores	**89**	*by K. G. Warren, B.Sc.*
Mathematical Tools	**95**	*by Computer*
Theory of Waveguides and Cavities	**98**	*by R. A. Waldron, M.A.(Cantab.)*
Measurement of Impedance at Audio-Frequency	**106**	*by N. P. Scholes and J. E. Macfarlane, B.Sc.(Eng.)*
Correspondence	**107**	
New Books	**109**	
Standard-Frequency Transmissions	**111**	
New Products	**112**	
Abstracts and References	**A37–A50**	

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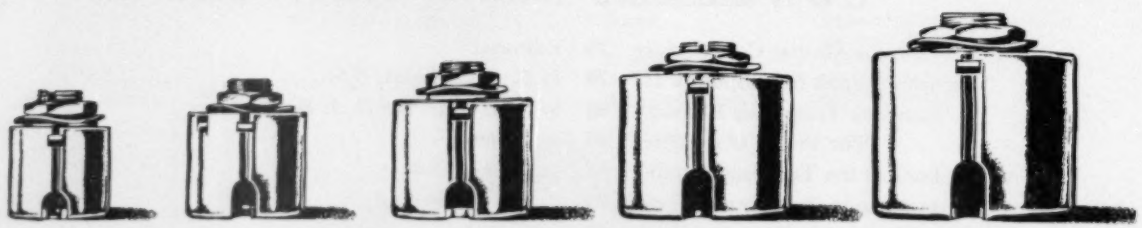

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Transistor Mutual Conductance

WE have on several occasions discussed the form of connection of the junction transistor and we have advocated that, as a matter of convention, the common-emitter connection should be regarded as the normal one for which the transistor constants are quoted and to which standard symbols apply. We have been stimulated to discuss this by the correspondence which has been going on for some time now and which still continues.

We pointed out last month that if the further step is taken of using mutual conductance as a transistor constant instead of current amplification factor the same equivalent circuit is applicable to both valve and transistor, the ordinary negative-grid valve being a special simple case.

In fact, mutual conductance is being increasingly used, so our suggestion is not as revolutionary as it may seem to some.

No one doubts that the valve is a voltage-operated device in the sense that the anode current is controlled by the grid voltage. With the negative-grid valve there is, ideally, no grid current and it is impossible to entertain any idea of current control.

It is usually considered that the transistor is a current-controlled device in the sense that it is the base current which is the operative factor in controlling the collector current. It is said that, physically, it is the current in the base region which is the determining factor rather than any electric field due to any externally-applied voltage.

Be that as it may, it is impossible to change the base current without also changing the voltage between base and emitter, since the transistor has a finite input resistance. Most people, indeed, would regard the voltage as the primary force and the current as the secondary effect.

Even if within the transistor it is current which affects current, therefore, it is quite a legitimate way of thinking to consider a change of collector current as being primarily caused by a change of base–emitter voltage.

In all matters of this kind very much is merely convention. One system may be fundamentally as good as another; one may be more convenient in some applications, the other may be better in others. The specialist is familiar with them all and finds no difficulty in switching between them as occasion arises. Others find it very confusing, especially students, and we think that there are great advantages to be gained by adopting a system which enables a uniform small-signal theory to be used for both valves and transistors.

Telemetry Signals From Sputnik III

INTERPRETATION AND TRANSCRIPTION OF DATA

By R. E. Henderson B.Sc.*

SUMMARY. *Equipment is described for transcribing telemetry signals from Sputnik III from magnetic tape on to photographic film. The resultant record on 35-mm film is in raster form showing successive keying cycles one under another. Changes in the pattern of pulses can thus be readily detected by eye and, by means of a time scale recorded on the edge of the film, the time they occurred can be read off. Sample records are shown. A description is reproduced of the telemetry encoding system used in the satellite and the results of the analysis of two transits are included and briefly discussed.*

When Sputnik III was first launched on 15th May 1958, the signals from it on a frequency of 20 Mc/s consisted of groups of four pulses repeated cyclically. Each group lasted 1·23 sec and comprised a marker pulse of 300 msec duration, followed by three information pulses. It is believed that the first of these indicated by its duration whether solar or chemical batteries were being used. The second and third information pulses gave cosmic-ray data by changes in their duration. After about a month of operation, the marker and battery pulses ceased, leaving the two cosmic-ray pulses with altered characteristics and a repetition rate of twice the original[1].

This paper describes equipment which has been developed for the extraction of the cosmic-ray information from magnetic-tape recordings of these signals.

Principle of Operation of Equipment

It would be possible to obtain photographic records by passing a film horizontally in front of a cathode-ray tube screen on which the beam is deflected vertically by the signal but this has two disadvantages compared with the system adopted. It is wasteful of film and, since the information is given by changes in pulse length, each pulse in a group is inconveniently distant from the corresponding pulse in successive groups.

In the system used, the signal is displayed on a linear time-base having a sweep time of 1·23 sec, the same as the period of the telemetry coding. The film is passed across the tube in a direction normal to the time-base so producing a raster type of display.

By folding up the time scale along the film in this way, a saving of approximately ten to one in film length is obtained for the same effective scale length and, at the same time, comparison of successive groups of pulses is facilitated.

A double-beam oscilloscope, with the internal time-base disconnected, is used for the display. The externally-generated time-base waveform is applied directly to the Y_2 plate and the signal to the common X plate. The Y_1 beam is used to produce timing marks on the film at 10-second intervals. In the camera, a continuously-moving film traverses the tube screen in a direction parallel to the X-axis thus separating successive strokes of the time-base on the film.

* *D.S.I.R. Radio Research Station, Slough, Bucks.*

Circuit Description

Time-Base Generator

In Fig. 1, the time-base generator V_4 is a conventional Miller integrator-transitron. The voltage which supplies the integrating circuit is stabilized by V_3 to improve linearity and to ensure a constant repetition rate. Coarse and fine time-base speed controls VR_1 and VR_2 are used and the clamping diode V_5 gives control over the amplitude of the sweep and also ensures a rapid anode-voltage recovery. The flyback time is about 5 msec which represents a loss of only 0·5% in the time scale.

The time-base generator is followed by a directly-coupled amplifier V_6 which provides an output amplitude of 200 V peak to peak for deflection of the oscilloscope Y_2 beam. The output stage V_7 and V_8 provides direct coupling to the c.r.t., maintaining the wide sweep but converting the mean potential from 150 V to approximately earth potential. This is necessary to prevent unwanted shift of the Y_2 beam by the d.c. component of the output of V_6. Although the stabilizer V_7 is operated below its normal minimum current rating, tests show that the linearity is more than adequate.

Time-Mark Generator

A flip-flop circuit V_2 triggered from a source of timing pulses produces a pulse of two-seconds duration every ten seconds. The direct coupling to the Y_1 plate of the oscilloscope produces a steady shift which holds the trace clear of the signal beam. The time-mark pulse deflects the beam off the screen leaving a gap in the trace.

The recorded signal from the satellite is applied to the X-channel of the c.r.t. Since the X-deflection plate is common to both beams of the tube, there is some loss of timing accuracy on the film due to displacement of

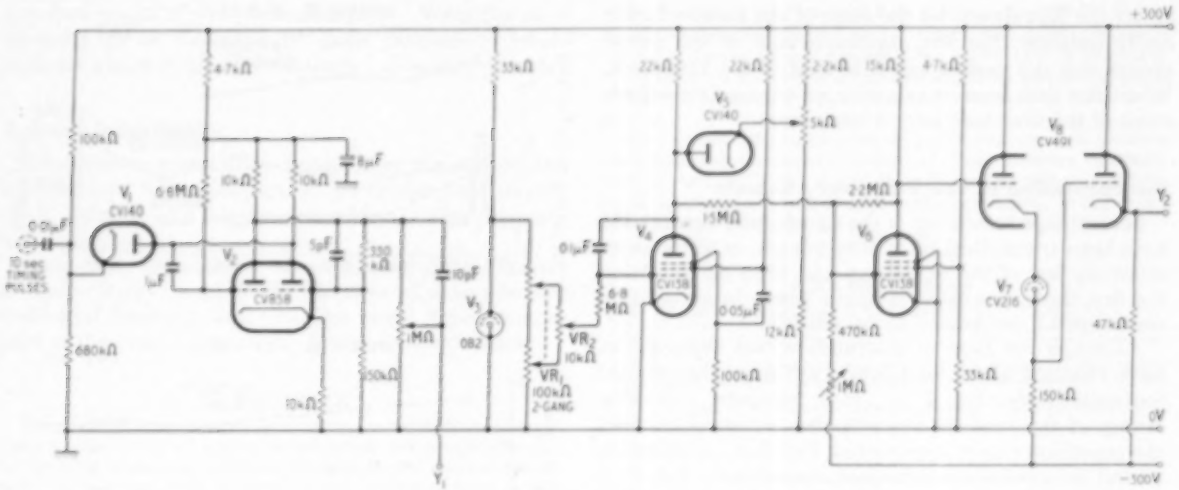


Fig. 1. *Circuit of time-base and timing unit*

the edges of the timing pulses by this signal. An alternative method of timing, suitable also for single-beam tubes, would be intensity modulation of the signal beam.

Signal Amplifier

A simple amplifier and gain-control are built into the equipment so that the level of the signal applied to the c.r.t. from the tape recorder may be maintained at the optimum value.

Photographic Arrangements

In the 35-mm camera used for photographing the c.r.t. trace, the linear reduction (object to image size) is 2·6 : 1 and the film speed 1·1 mm/sec. Since the duration of the time-base sweep is 1·23 sec, the separation between scans on the film is 1·35 mm and this permits a maximum peak-to-peak signal deflection on the tube of 3·5 mm if no overlapping is to occur. The length of the time-base scan is 5·85 cm which gives an image of approximately 2·3 cm on the film, leaving space for the timing trace along the edge of the film.

Operation

For convenience, a second oscilloscope is used to monitor the signal level, its X-plate being connected to the 1·23-sec time-base and its Y-plates to the signal output.

The time-base period is adjusted by taking the output to the gate of an electronic counter which is fed with a standard frequency. Thus, if the standard frequency is 1 000 c/s, the coarse and fine sweep-controls are adjusted to give a count of 1,230 per period of the time-base.

Some typical film records obtained with the equipment are shown in Fig. 2. The slow drift of the pulses

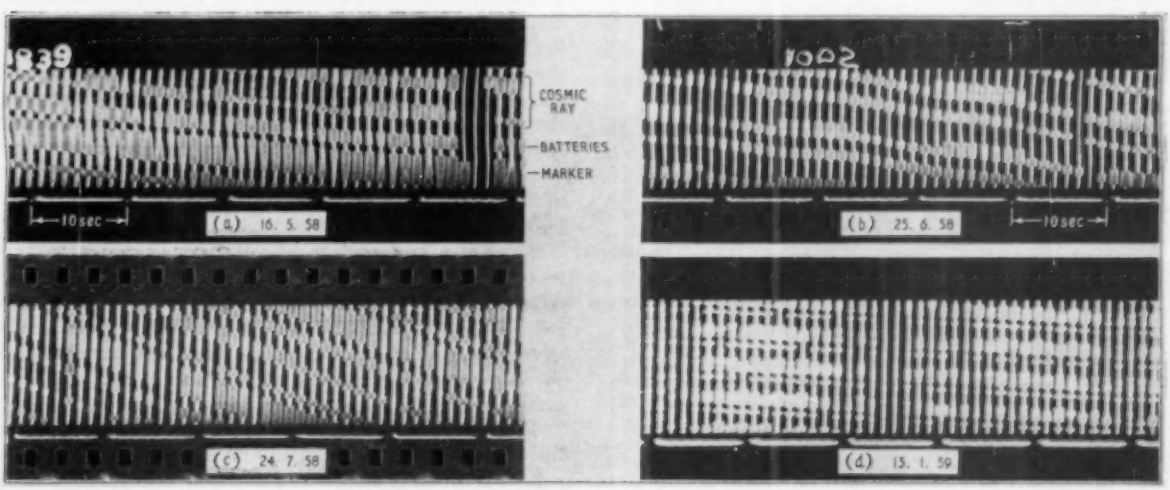


Fig. 2. *Sample film records*

across the film shown by the slope of the pattern in the figure indicate that the repetition rate of the signal groups was not exactly equal to that of the time base. When this drift became excessive an appropriate adjustment of the time-base period was made.

Interpretation of the Telemetry Signals

Several tape recordings of the signals from Sputnik III have been transcribed on to film though, unfortunately, relatively few of the recordings had been made during the first three months of its flight when the data being transmitted were known to be reliable.

Although the type of transmission was reported[1] to have changed about mid-June 1958 from the original four-pulse code, Fig. 2 (a), to a two-pulse code consisting of the two cosmic-ray pulses repeated at twice the repetition rate of the original, Fig. 2 (b), a transition period occurred when either was transmitted. Fig. 2 (c) shows a section of four-pulse code at that time. The pulses appear to be slipping, the marker appearing where an information pulse would normally be expected. By this time also, the section of the satellite equipment transmitting data on the rate of incidence of cosmic-ray particles had ceased operation[1,2]. After about two months of the second type of transmission a further change in the system was evident since the pulse configuration changed spasmodically and was not according to the published details[1], [Fig. 2 (d)].

Further particulars about the encoding arrangements in the satellite are given in the Appendix. It may be noted here that three items of cosmic-ray information were transmitted by the third and fourth pulses of the four-pulse code[1]. Each of these pulses could take on one of three durations, nominally 50, 100 and 150 msec, and the information was transmitted by commutation between these three pulse-lengths. Thus, if the duration of the third pulse remained at 50 msec for a period of T sec or departed from 50-msec duration for an interval of T sec, the mean particle flux intensity was $2 \times 10^9/T$ eV/sec over the time T. Similarly for the fourth pulse, the relation between mean intensity and T being $18 \times 10^9/T$ eV/sec.

The number of particles incident was determined by dissection of the time intervals during which the duration of either the third or fourth pulses departed from 50 msec (the second of the alternatives mentioned in the preceding paragraph). During these sub-intervals, the pulse length was either 100 or 150 msec. It will be clear from the explanation in the Appendix that the two terminal sub-intervals have to be ignored.

If T' seconds was the time between the changes of the third or fourth pulse duration from 100 to 150 msec or vice-versa, the mean rate of incidence of particles detected by the counter in the satellite during T' was $2\,048/T'$ per sec, since 2 048 particles were counted in this interval.

Examples of Results

Typical graphs of the cosmic-ray data obtained in this way are shown in Fig. 3. Fig. 3 (a) shows the analysis of a recording made while the satellite was moving south-east from approximately 100 miles south of

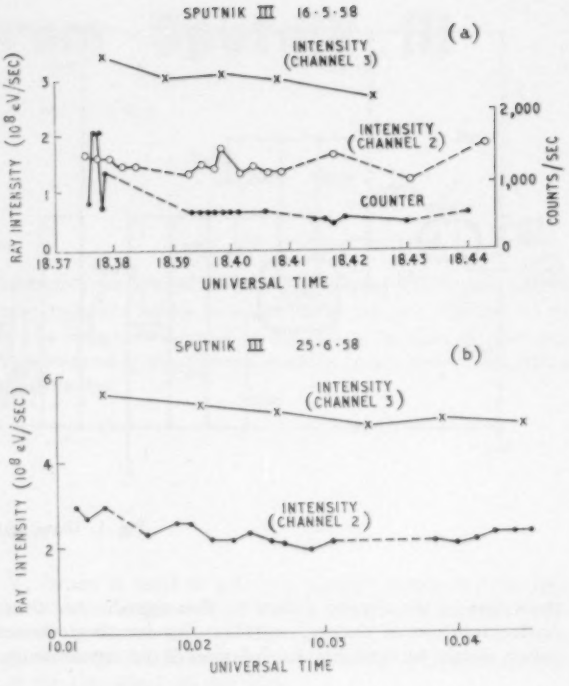


Fig. 3. *Sample graphs of cosmic-ray data*

Iceland to Corsica at a mean height of 600 km on the 16th May 1958. Fig. 3 (b) shows results from a recording on the 25th June 1958 with the two-pulse code only. The track in this case was again south-east, from Southern Ireland to North Africa at a mean height of 850 km.

Since the indicated time between pulse-duration changes is an integral number of 1·23-sec periods the possible values for the indicated rate of counting are in discrete steps which became further apart, with correspondingly poorer accuracy, as the count rate approaches its apparent limit. This occurs when the pulse-duration changes every successive group, that is T' is 1·23 sec and the count is 1 666 particles per second. At higher rates, the indicated count rate is dependent on the actual count rate and the pulse-group repetition rate; that is, the sampling rate.

In Fig. 3 (a) the system appears to be working at about its maximum rate of 1 666 counts per sec and there is the possibility of the system being unable to cope with the rate of information.

In only a few instances was there evidence on our records of overloading of the telemetry system on the intensity channel carried by the third pulse. Australian records, however, have shown practically continuous overloading on this channel[2]. This they attribute to the passage of the satellite through the inner radiation belt[3].

The discrepancy between the values of intensity given by the second and third channels indicates that the energy released in the crystal of the detection instrument (see Appendix) is by high-energy particles producing flashes of light strong enough to cause non-linearity in

the final stages of the photomultiplier. Were the same intensity to be indicated by both channels it would indicate a high number of relatively low-energy particles.

Acknowledgements

The author gratefully acknowledges the advice and assistance of his colleagues, in particular that of Mr. G. E. Ashwell who suggested the form of film presentation described.

This paper is published by permission of the Director of Radio Research of the Department of Scientific and Industrial Research and describes work carried out as part of the programme of the Radio Research Board.

APPENDIX

The detection instrument in the satellite was a scintillation counter using a large crystal of sodium iodide which had a high efficiency for detecting low-energy photons generated by the incident cosmic-rays. It was thus able to detect readily incident electrons with energies of 100 keV.

Three items of cosmic-ray information were transmitted by the third and fourth pulses in the four-pulse code. Details of the way these pulses are produced have been given elsewhere[1,2]. It will be sufficient to note here that the pulse lengths in each of the two channels are controlled by the positions of three single-pole change-over switches A, B and C in the satellite. The tables below show the nominal pulse-length in milliseconds transmitted for the various possible positions of the switches.

Length of Third Pulse (referred to as Channel 2)
(Switch C does not operate on this channel)

Switch B	Switch A	
	Position 1	Position 2
Position 1	50 msec	100 msec
Position 2	50 msec	150 msec

Length of Fourth Pulse (referred to as Channel 3)
(Switch A does not operate on this channel)

Switch B	Switch C	
	Position 1	Position 2
Position 1	150 msec	50 msec
Position 2	100 msec	50 msec

The state of these switches is examined every scanning cycle, that is, every 1·23 sec. Now switches A and C are controlled respectively by the integrated anode current of a photomultiplier and the integrated current from the seventh dynode of the multiplier. Each change in position of switch A represents an increment of 2×10^9 eV in the energy lost by the cosmic-rays in the sodium-iodide crystal, the scintillations in which actuate the photomultiplier[1]. On the other hand, each change in switch C corresponds to 18×10^9 eV, and the switch accordingly changes over less frequently than does A.

The position of switch B is controlled by a scaler, the input of which is connected to the last dynode of the photomultiplier. The constants of the scaler are such that a change in position of switch B occurs for every 2 048 scintillations produced by the incident particles.

Examination of the tables in conjunction with the above description leads to the condensed statement of the operating characteristics of the telemetry channels given in the text. In addition, it will be noted from the tables that when switch A is in position 1 the length of the third pulse is 50 msec for both positions of switch B; that is, this channel cannot transmit particle-count information. Similarly, the other channel cannot respond when switch C is in position 2. When these conditions exist simultaneously, as occasionally happens, no particle-counts are transmitted and such periods have to be ignored.

Other complications arise when the switches change their positions more than once during the interval (1·23sec) between successive scans of their positions. A full discussion of this difficulty will be found in reference 2; the illustrative examples given in Fig. 3 are not vitiated by these effects.

REFERENCES

[1] S. N. Vernov, A. E. Chudakov, E. V. Gorchakov, Ju. I. Logachev and P. V. Vakulov, "Study of Cosmic-Ray Soft Component by the Third Soviet Earth Satellite", Artificial Earth Satellites, No. 2, p. 61, USSR Academy of Science, Moscow 1958.

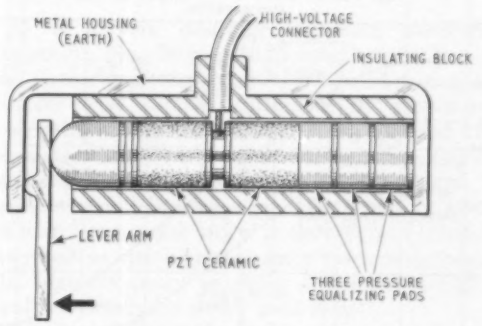
[2] A. J. Herz, K. W. Ogilvie, J. Olley and R. B. White, "Radiation Observations with Satellite 1958 δ over Australia", *Nature, Lond.*, 8th Aug. 1959, Vol. 184, p. 391.

[3] J. A. Van Allen, C. E. McIlwain and G. H. Ludwig, "Radiation Observations with Satellite 1958 ε", *J. geophys. Res.*, 1959, Vol. 64, p. 271.

NEW PIEZOELECTRIC IGNITION SYSTEM

Recent developments in piezoelectric materials have enabled high-voltage impulse sources to be built based on the direct conversion of mechanical energy to electrical energy. The principle of operation has been exploited by the Clevite Corporation, U.S.A., in their 'spark pump' as an ignition device for internal combustion engines, and their associate company Brush Crystal in England are extending the principle for other applications.

The Clevite 'spark pump' consists of two cylinders of PZT* ceramic enclosed in a plastic housing and provided with a lever arm which is designed to be driven by an eccentric on the engine crankshaft or camshaft (see figure). When driven, the lever arm exerts



A sketch showing the basic principles of construction of the 'spark pump'

an axial pressure on the ceramic elements and the generated voltage developed across the elements is sufficient to produce an ignition-plug spark. The timing switch to control the spark can be operated by the flywheel.

The ceramic elements are 0·75 in. long by 0·375 in. diameter and with a lever multiplication of 15 : 1 a static force of 80 lb will produce voltages in the range 16,000 to 21,000 V. The actual force applied to the ceramic cylinders is in the region of 7,000 p.s.i. This is below the safe stress limits of the ceramic elements which can be stressed to provide voltages up to 30,000 V.

In order to achieve a reasonable charge energy, the two ceramic cylinders can be connected in parallel and, with stresses of the order quoted, the operating life of the ceramic elements is claimed to be in excess of 10^7 cycles of compression.

* A registered trade name for the zirconate-titanate group of ceramics developed by Clevite Corporation.

Transistor Frequency Response

GRAPHICAL PRESENTATION OF f_T

By J. R. James* and D. J. Bradley*

SUMMARY. *Recent transistor specifications usually include typical values of f_T. This article describes a graphical presentation of f_T as a function of the d.c. working point and the advantages of this method.*

In this article the importance of f_T as a transistor figure of merit is discussed, f_T being the frequency at which the small-signal short-circuit current gain in the common-emitter (c.e.) configuration is unity. Before proceeding further, it is of interest to reflect on the existing methods of specifying transistors in order to show that an improved means of specification would be welcome.

A device manufacturer often seeks simplicity when deciding on a test specification, while the user desires the maximum amount of information for design purposes. Both lines of approach are influenced by economics.

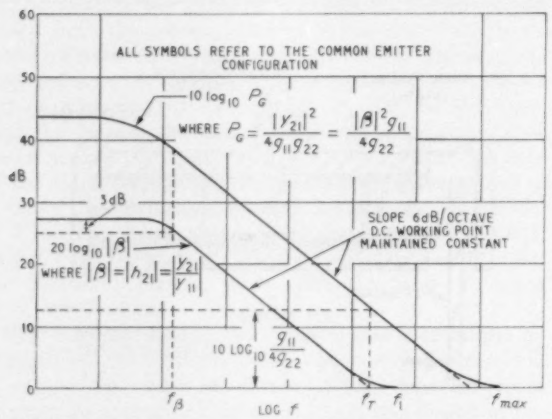


Fig. 1. *Frequency response in the common-emitter configuration*

Complications occur with transistor specifications because the device is not fully understood. For instance, it was shown[1] recently that the c.e. configuration is the basic mode of operation in a transistor and, therefore, the device should be specified in terms of its c.e. parameters. Also, despite duality theories, linear transistor circuits are usually analogous to their valve counterparts and, to some extent, this analogy is true of non-linear applications. While it may seem logical to be guided by the well-established methods of specifying valves, the inferior frequency response of transistors must be defined in any specification.

The hybrid-π equivalent circuit and the matched unilateralized power gain (shown by Mason[2] to be a mathematical invariant) are two accepted methods of describing the frequency response of transistors. The disadvantages associated with these methods are well known; both are difficult to measure and are only true for a predetermined working point, while the hybrid-π presents the designer with cumbersome calculations.

In comparison, the graphical presentation of f_T, which is fairly simple to measure, embraces a variety of working points and, due to its simplicity, promises to be an excellent means of describing the high-frequency behaviour of transistors.

The examination of the frequency response in the c.e. configuration (basic mode of operation) yields some basic relationships concerning f_T (Appendix 1). Typical experimental curves of maximum matched unilateralized power gain P_G and $|\beta|$ are shown in Fig. 1. P_G is unity at f_{max}, while $|\beta|$ is unity at f_1 and, at f_β, it is 3 dB down on its d.c. value. The projected value of f_1 is given by f_T. The latter differs very little from f_1 and, by extrapolation, f_T is much easier to measure; therefore, it is generally accepted in place of f_1.

Typical experimental plots of f_T as a function of the working point $[f_T\ (V, I)]$ are shown in Figs. 2 and 3 for a high-frequency amplifier and switch respectively. The differing characteristics of these two types are readily shown by the position of the 'ridge' in the f_T contours.

Estimation of Circuit Performance Using f_T (V,I) Curves

In most cases the formulae used in this article are approximate. It is intended to anticipate only the basic rules used by experienced designers when initiating a design. The ease with which these 'rules of thumb' are applicable to the f_T $(V,I.)$ curves will now be discussed.

(a) Narrow-Band Tuned Amplifiers

Using the f_T (V,I) curves a designer can select an operating point P_0, Fig. 2, which gives the maximum matched unilateralized power gain P_G

* *Semiconductors Ltd.*

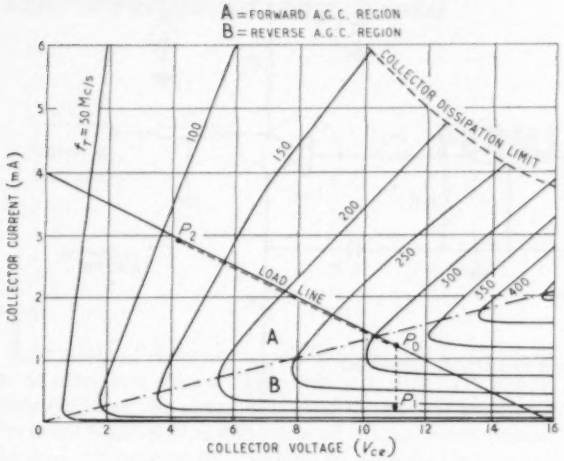


Fig. 2. f_T (V, I) *plot of typical amplifier transistor*

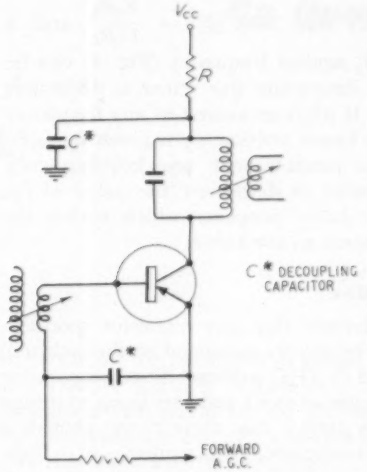


Fig. 4. *Tuned amplifier using forward a.g.c.*

for a minimum current consumption from a given battery. It is shown in Appendix 1 that P_G is approximately proportional to f_T squared and the point of maximum P_G and f_T coincide.

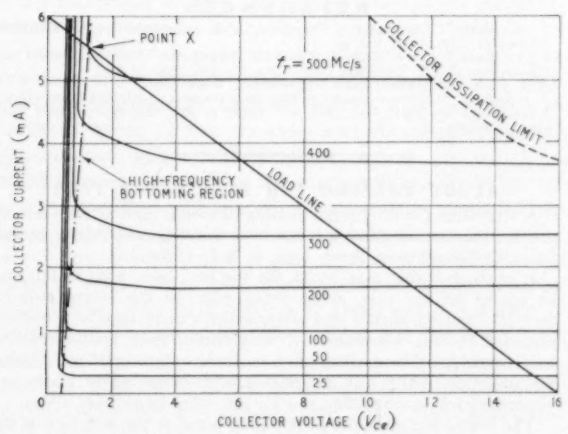
The reduction in gain under a.g.c. action can be estimated by noting the change in f_T. The error in this estimation is obviously increased by mismatch losses and detuning effects. The movement of P_0 to P_1 is under normal (reverse) a.g.c. while P_0 to P_2 corresponds to forward a.g.c. The basic circuit for forward a.g.c. is shown in Fig. 4. The polarity of the diode producing the a.g.c. error signal is such that the collector current increases as the signal level and the working point moves along the load line. The two regions of reverse and forward a.g.c. that are positioned on opposite sides of the 'ridge' are shown in Fig. 2.

(b) Video Amplifiers

The significance of f_T as a figure of merit for video amplifiers is well known and briefly described in Appendix 2. Using the f_T (V,I) curves, the designer can select the best f_T for a given battery voltage. The large-signal f_T will of course differ from the low-signal one but there is a correspondence between them; i.e., they increase and decrease in step.

(c) Switching Circuits

f_T is a very desirable switching parameter (Appendix 3). By superimposing a load line on Fig. 3, the approximate f_T experienced during the turn-on cycle can be computed by graphically integrating the f_T encountered along the load line. Beyond point X on the load line f_T drops rapidly and an increasingly large drive is required to maintain a fast leading edge. The point X where the load line cuts the 'ridge' in the contours could be described as the high-frequency bottoming point. The high-frequency amplifier transistor of Fig. 2 would obviously be a very poor switch.



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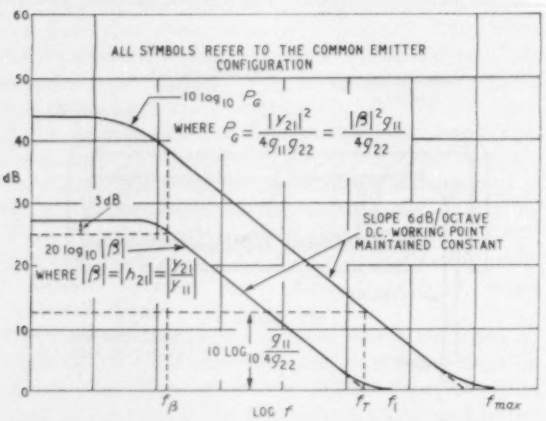

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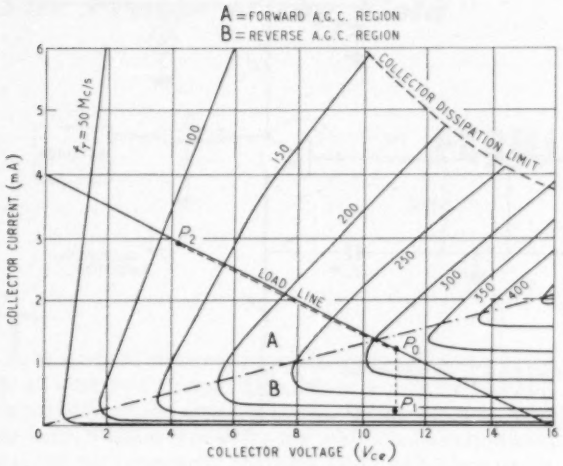

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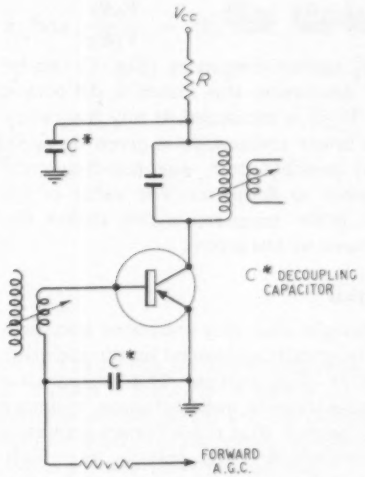

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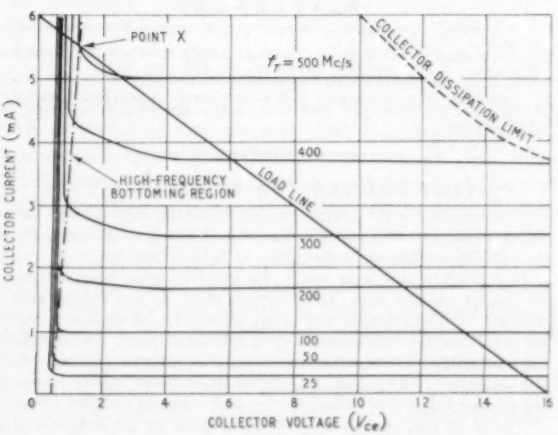

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Once set up, the equipment is very easy to operate.

It is easily seen that $|\beta| = \frac{V_0 R_1}{V_1 R_L}$ and a plot of $|\beta|$ against frequency (Fig. 1) can be made to determine the linear 6 dB/octave portion. If $|\beta|$ is measured at any frequency f_0 on this linear portion, f_T is given by $f_0|\beta|$. It is thus possible with one low-frequency measurement to determine the value of f_T. It is this latter property which makes the measurement so attractive.

Conclusion

It is thought that any transistor specification will be greatly enhanced by the inclusion of typical f_T (V,I) curves. The comparative ease of measurement and usefulness to design engineers suggest that these curves provide a most satisfactory description of high-frequency behaviour, from economic considerations.

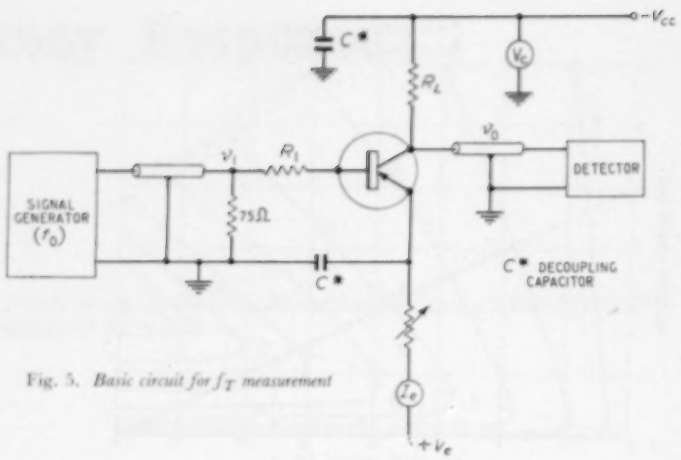


Fig. 3. *Basic circuit for f_T measurement*

Acknowledgements

The authors are indebted to the management of Semiconductors Limited for permission to publish this article. Thanks are due to Mr. J. W. Ringrose, Head of the Applications Laboratory, who conceived the usefulness of the f_T (V,I) curves and encouraged the writing of this article.

APPENDIX 1

Details of the matrix parameters used in this article are given in the literature[3]. y parameters are basic to the c.e. configuration and are therefore used to express the maximum matched unilateralized power gain P_G.

$$P_G \simeq \frac{|y_{21}|^2}{4 g_{11} g_{22}} = \frac{|h_{21}|^2 g_{11}}{4 g_{22}} = \frac{|\beta|^2 g_{11}}{4 g_{22}} \quad .. \quad .. \quad .. \quad .. \quad (1)$$

In Fig. 1 the curves of $|\beta|$ and P_G have the same slope of 6 dB/octave between f_β and f_T and the following relationships apply to this region. Taking decibel values of Equ. (1),

$$10 \log_{10} P_G - 20 \log_{10} |\beta| = 10 \log_{10} \frac{g_{11}}{4 g_{22}} = \text{constant (between } f_\beta \text{ and } f_T)$$

Therefore, at a given d.c. working point g_{11}/g_{22} is a constant independent of frequency between f_β and f_T. Two other relationships can now be easily derived in terms of f_0 the operating frequency.

$$P_G = \left(\frac{f_{max}}{f_0}\right)^2 \Bigg|_{f_\beta < f_0 < f_{max}} \quad .. \quad .. \quad (2)$$

$$P_G = \left(\frac{f_T}{f_0}\right)^2 \frac{g_{11}}{4 g_{22}} \propto (f_T)^2 \Bigg|_{f_0 = \text{constant}} \quad .. \quad .. \quad (3)$$

Equ. (2) is useful at a fixed d.c. working point while (3) can be used to estimate the gain at any d.c. working point. It will be found in practice that the term $\frac{g_{11}}{4 g_{22}}$ is fairly constant in the forward a.g.c. region and more variable in the reverse a.g.c. region. However, this term is still a slow moving function compared with $\left(\frac{f_T}{f_0}\right)^2$. If P_G is specified at a given d.c. working point then, using the corresponding f_T, the term $\frac{g_{11}}{4 g_{22}}$ (Fig. 1) is defined. This gives the turns-ratio of the matching transformer.

While the comments in this Appendix depend to some extent on the type of transistor in use, Equ. (3) is a very useful approximation.

APPENDIX 2

Video amplifiers are not matched but terminated by the input admittance of the following stage. The term 'iterative' is associated with this mode of operation. The effect of the mismatch on the gain P_{GV} can be estimated as follows (where all symbols refer to the c.e. configuration).

$$P_{GV} = P_G - \text{(mismatch losses)}$$
$$= \frac{|y_{21}|^2}{4 y_{11} y_{22}} \div \frac{(y_{11} + y_{22})^2}{4 y_{11} y_{22}} \simeq \left|\frac{y_{21}}{y_{11}}\right|^2 = |\beta|^2$$

$|\beta|$ = voltage gain = current gain (since P_G = voltage gain × current gain). Voltage gain × bandwidth = $|\beta_c| f_\beta = f_T$ where $|\beta_c|$ is the $|\beta|$ at f_β.

APPENDIX 3

Transistor switching circuits usually employ a capacitor drive to the base, in order to achieve a fast rise time with minimum hole storage. The charge Q_{on} that must be delivered by the capacitor is given by the following equation[4].

$$Q_{on} = \frac{I_c}{2\pi f_T \alpha_0} + C_{tc} \Delta V_{cb} M$$

I_c = the collector current

α_0 = small-signal common-base short-circuit current gain of the transistor when $V_{cb} = 0$

M = a factor between 1 and 2 for alloy junction transistors and between 1 and 1·5 for grown junction transistors

ΔV_{cb} = change in collector voltage

C_{tc} = the capacitance at the 'off' collector voltage

f_T varies considerably during the turn-on cycle and the mean f_T can be estimated using the load line in Fig. 3.

REFERENCES

[1] J. R. James, "The Junction Transistor". A communication to *Electronic Technology*, Sept. 1960, Vol. 37.

[2] S. J. Mason, "Power Gain in Feedback Amplifiers", Research Laboratory Electronics, M.I.T., Technical Report No. 25, Jan. 1953, p. 14.

[3] R. F. Shea, "Transistor Circuit Engineering", Wiley 1957.

[4] J. J. Sparkes, "The Measurement of Transistor Transient Switching Parameters", I.E.E. paper No. 3006, Jan. 1960, Vol. 106B, p. 562. "I.E.E. Transistor Convention 1959".

ALLOY COATING FOR TRANSISTOR TABS

A technique for the partial coating of metals used as base tabs in the manufacture of transistors has been developed by Alpha Metals Inc. (56 Water Street, Jersey City, 4, N.J., U.S.A.).

It is claimed that as a result, for the first time, base metals can be coated on one side, along either edge, in the centre only or over-all. Typical metals and alloys which can be used as a coating material are tin, tin-antimony, tin-gallium, lead, lead-antimony, tin-lead-gold, tin-lead, silver, indium, and indium gallium. Along the materials which can be coated with these solder alloys are copper, nickel, kovar, rodar, therlo, and other iron-nickel alloys.

The technique produces a true alloy bond at the interface of the two metals, assuring a low-resistance junction and good ohmic contact.

FERMI LEVEL

The study of electrons in solids really has to be taken in at least two stages. The first is a relatively simple elaboration of the classical kinetic theory of gases, with the usual idealizations and a few additional assumptions; this has not necessarily anything to do with electrons or with solids, and would be equally applicable to, say, the right sort of helium atoms—the question of electric charge does not come into it. In the second, we have to take into account both the electron charge, and also the periodic electric field of the lattice of the solid; and this is where things do start to become difficult, for it is like the business of adapting the kinetic theory of the ideal gas to meet the case of a real gas which has all sorts of distinctive physical properties to reckon with. But, providing that we take things steadily and bear this analogy with the kinetic theory in mind, it should at least be possible to explore the general ideas that are used.

In the classical kinetic theory of the ideal gas, the individual particles possess mass, momentum, and kinetic energy—and not much else. The only other important quantity is the number of particles per unit volume. We have to suppose that they have appreciable though very small dimensions, so that they can reach equilibrium conditions by colliding with one another, but these dimensions do not at this stage need to be specified (mean-free-paths, and so on, belong to *real* gases). The pressure exerted by the gas on the walls is found in terms of the *momentum* change. The absolute temperature, under equilibrium conditions, is proportional to the average kinetic energy; and the total internal energy per gram is $U = 1{\cdot}5\,RT/M$, where M is the molecular weight and R the universal gas constant, whence the specific heat, dU/dT, equals $1{\cdot}5\,R/M$.

In order to calculate the distribution of the energies between the particles (that is, the number dN with energies between W and $W + dW$) it is necessary to work in terms of the *momenta*, and to construct a six-dimensional space called phase space, in which a particle is represented by the three-position co-ordinates x, y, z, and the corresponding momentum co-ordinates p_x, p_y, p_z. This space is divided up into elementary cells, which are *unrestricted as to size*, and *unrestricted as to the values of p;* I nearly said, unrestricted as to the number of particles that can occupy a given cell—but this is proportional to the size of the cell, and is the key to the whole of the rest of the calculation. For, from the theorem that the number of occupants is proportional to the size of the phase-space cell, the Maxwell-Boltzmann distribution of *energies* is obtained. I need not go further here into Maxwell-Boltzmann statistics—this is worked out in all the books (for example, G. Joos' "Theoretical Physics", Chapter XXXIV).

For the three important points are the difference that it makes to the distribution of energies when we introduce Planck's constant h. For then we find that

(1) the values of p are confined to integral multiples of $h/2\pi$;

(2) the minimum volume of an element in phase space is h^3, as the Uncertainty Principle shows, since the uncertainty Δx in a position co-ordinate and the corresponding uncertainty Δp_x in the momentum are related by $\Delta x \,.\, \Delta p_x = h$; and

(3) not more than one particle (or, for electrons with spins of opposite signs, two particles) can occupy the same cell, by the Exclusion Principle.

Maxwell-Boltzmann Statistics and Quantum Statistics

The Uncertainty Principle and the Exclusion Principle are not just arbitrary rules, any more than is the business of quantizing momentum and also energy into finite discrete values. They can be justified by quantum-mechanical theory which, in fact, had to adjust itself to cope with Uncertainty and Exclusion as they emerged.

The choice of an appropriate statistical treatment is also not arbitrary, for it depends on the way in which the particles react to one another. The term "weak interaction" is used in Pauling and Wilson's "Quantum Mechanics" for the process by which a particle is able to divine the nature of the other particles round it and behave appropriately; and in that book, the onus is shifted on to the form of the wave-function that describes the particle. This may be symmetrical (indicating a spin of 0 or ± 1); or anti-symmetrical (spin of $\pm\frac{1}{2}$); or both. In crude pictorial terms, this can be taken as indicating that the particles tend to attract one another (not in the mechanical sense, but tending to crowd into the same energy states); or to avoid one another, so that the occupation of an energy state by one particle means that it is not available to others; or that they have no influence on one another's actions at all.

In the last of these cases only do we get completely random unbiased chance or chaos operating; this is the condition for a Maxwell-Boltzmann distribution of energies, in which the only thing that determines the probability (for any given equilibrium temperature T°K) of a particle possessing a chosen energy is the magnitude of the energy itself. In *classical* mechanics, the probability $P\,(W)\,dW$ of the energy of a particle

lying between W and $W + dW$ is given by an expression containing the Boltzmann exponential factor $e^{-W/kT}$. In quantum mechanics, this has to include as a factor $p(W)$, where $p(W)dW$ is the number of possible energy states lying within this range, and that is the only modification.

But, if the particles tend to crowd into states occupied by others, the exponential term is replaced by the factor $(A\,e^{W/kT} - N)^{-1}$; here A is a constant chosen so as to make the sum of all the probabilities for all the states equal to unity, and N is the total number of particles. This result is obtained by Bose-Einstein statistics.

And if the Exclusion Principle operates, the exponential factor is replaced by $(A\,e^{W/kT} + N)^{-1}$. This is the Fermi-Dirac result, which applies to electrons.

However, in a number of cases it may happen that $Ae^{W/kT}$ is very large compared with N; in such cases the other two factors both approximate to $Ae^{-W/kT}$, so that we get the Maxwell-Boltzmann distribution as a limiting case.

The Degenerate Electron Gas

As we are going to discuss the Fermi-Dirac distribution for a gas of free electrons in a metal, the first point to settle is why the Maxwell-Boltzmann one won't do. In terms of (3) above, one answer can be found in terms of the wave-functions, which are anti-symmetrical for electrons; or, more simply, in the fact that the electron spin is $\pm\frac{1}{2}$. Both these statements amount to the same thing, and mean that the individual particles can be distinguished from one another. In terms of the specific heat, there should be a very large contribution $1{\cdot}5\ R/M$ (the molecular weight of electron-gas being about 1/2000!) so that all metals should have huge specific heats, which they don't. Since the electron-gas is apparently refusing to play, we blacken its character and say that it is *degenerate;* but the real fault lies with the Maxwell-Boltzmann distribution which is refusing to work. There are signs that, at a sufficiently high temperature, some electrons do in fact make some contribution to the specific heat, and have become undegenerate (or regenerate, if you like); but, as will appear later, this temperature has to be very high indeed. And at all temperatures they make the more limited contribution allowed by Fermi-Dirac statistics.

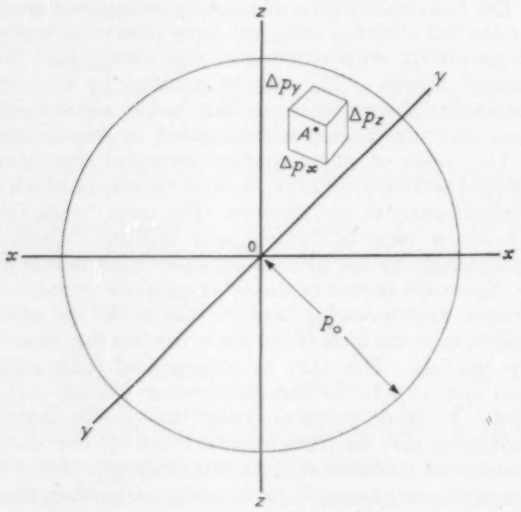


Fig. 1. *Momentum-space diagram. The co-ordinates representing a particle in three dimensions are the components p_x, p_y, and p_z of its momentum along the three axes. The momentum p of a particle A is thus represented by the point A on the diagram. But, by the Uncertainty Principle, this point can only be located to within the volume of momentum space represented by the little cube $\Delta p_x \cdot \Delta p_y \cdot \Delta p_z$. Each particle has this volume of momentum space allotted to it and, as the Exclusion Principle operates, no other particle can trespass on this volume. Taking 'particle' to mean a pair of electrons with opposite spins, we can see that for N electrons we need at least $N/2$ such little volumes; the least volume of momentum space that will accommodate them, and at the same time satisfy minimum-energy conditions, is a sphere of radius p_0*

The next point is, how many electrons ought we to consider? It is usual to suppose that each atom of a metal contributes one conduction electron; we shall imagine that it has done so and then vanished. Since different metals have different atomic weights and densities, the number of electrons per unit volume will depend on the actual metal that we start with. The lattice of positive ions is, of course, needed to restrain the electrons within the sample; and if we go and remove that, we shall have to remove the charges on the electrons as well. So that we are left with an assemblage of particles which have only mass, momentum, and kinetic energy, and can occupy any position within the boundary walls. If there are N atoms in a specimen of volume V, they give us an electron-gas comprising N electrons occupying a volume V.

Finally, the temperature. The degenerate electron-gas has no appreciable traffic with heat energy at ordinary temperatures; it begins to lose this character at much higher temperatures, so we must go the other way. We consider the gas to be at absolute zero. This always seems to be a difficult idea; first, because we believe that absolute zero is unattainable and this looks like a deliberate choice of ground remote from reality and, secondly, because the Boltzmann association of energy with kT is so thoroughly ingrained that it is hard to believe that there will be any energy to distribute. There wouldn't be with Maxwell-Boltzmann statistics, but these we know are not working. The real reason why we choose absolute zero is that it will give us the *least-energy* distribution, and that is what we want to find.

The Fermi Level

Consider N electrons contained in a volume V. If m is the mass of each, and p the momentum, then calling the velocity v, as $p = mv$, and the kinetic energy $W = \frac{1}{2}mv^2$, $W = p^2/2m$.

We can write down at once the total number of discrete energy levels. The Exclusion Principle, taking into account the two spins, shows that $N/2$ levels will be needed to accommodate all the electrons. We know that the lowest possible level has $p = 0$ and $W = 0$; where is the highest level for the least-energy conditions?

The first step is to use the Uncertainty Principle to work out the least separation of the levels. Suppose that this is achieved when the highest has a momentum p_0 and energy $p_0^2/2m$.

Working in phase space, of which only the momentum

co-ordinates are shown in Fig. 1 (which really represents *momentum* space), let x, y, z be the position co-ordinates and p_x, p_y, p_z the momentum co-ordinates as before. For each direction we have, for the uncertainties in position and momentum,

$$\Delta x \,.\, \Delta p_x = h = \Delta y \,.\, \Delta p_y = \Delta z \,.\, \Delta p_z,$$

so $\Delta x \,.\, \Delta y \,.\, \Delta z \,.\, \Delta p_x \,.\, \Delta p_y \,.\, \Delta p_z = h^3$.

But the chosen electron may be anywhere within the volume V, so

$$\Delta x \,.\, \Delta y \,.\, \Delta z = V, \text{ whence } \Delta p_x \,.\, \Delta p_y \,.\, \Delta p_z = h^3/V.$$

This is the range of momentum-co-ordinates within which an electron can be found. The whole of the momentum-space of Fig. 1 can thus be divided up into cells of this volume, each of which accommodates *two* electrons (Fig. 2). If p_0 is the *least value of the momentum for the highest level* which will accommodate the lot, the corresponding volume in momentum-space is $\frac{4}{3}\pi p_0^3$

Hence $\frac{4}{3}\pi p_0^3 = \frac{1}{2} Nh^3/V$,

$$p_0 = h\left(\frac{3}{8\pi}\frac{N}{V}\right)^{\frac{1}{3}}$$

and $W_0 = p_0^2/2m = \dfrac{h^2}{2m}\left(\dfrac{3N}{8\pi V}\right)^{\frac{2}{3}}$

W_0 is called the Fermi Level, and will be given the symbol W_F from now on.

The value of W_F is readily found for a metal of atomic weight A, since N/V is the same as the number of *atoms* per unit volume, which is $\rho/m_H A$, where ρ is the density and m_H the mass of the hydrogen atom. Putting in the numerical values $h = 6{\cdot}55 \times 10^{-27}$ erg-sec, $m = 9 \times 10^{-28}$g, and $m_H = 1{\cdot}66 \times 10^{-24}$g, we get

$$W_F = 4{\cdot}12 \times 10^{-11}\,(\rho/A)^{\frac{2}{3}} \text{ erg} = 25\,(\rho/A)^{\frac{2}{3}} \text{ eV}$$

in round figures, since 1 eV is equivalent to $1{\cdot}6 \times 10^{-12}$ erg.

We find, for example,

Metal	A	ρ(g/cm³)	W_F
Sodium	23	0·97	3·1 eV
Copper	63·5	8·9	6·6 eV
Silver	108	10·5	5·4 eV

Now, these energies are huge according to Boltzmann kinetic-theory ideas. For 1 eV corresponds to a kT temperature of about 10^4 °K, so that we have the extraordinary-looking situation of the electrons in a metal at absolute zero having energies of the order associated with something like 10^5 °K. But if we come to invidious comparisons, the mass of an electron itself on the mc^2 count is about 5×10^5 eV which we could, I suppose, write down if we wanted to as 3×10^{17} °K; this affronts common sense, as it cannot be related to ordinary thermal agitation—but neither can W_F at absolute zero.

The Fermi Distribution at Absolute Zero

The number of electrons dN with momenta lying between p and $p + dp$ can be found by dividing the volume of momentum-space within which the representative point must lie by h^3/V; thus,

$$dN = 4\pi p^2 dp\, V/h^3$$

Remembering that the energy W corresponding to momentum p is $p^2/2m$, and using the expression for W_F above, we get

$$\frac{dN}{dW} = \frac{3}{2}\frac{NW^{\frac{1}{2}}}{W_F^{3/2}}$$

The distribution curve is then of the form shown in Fig. 3, where it must be remembered that, although a smooth curve is drawn, the level structure of Fig. 2 is still there and it is not a *continuous* distribution.

Since the form of the distribution is known, the average energy can be calculated; this works out to 0·6 W_F, and this is the figure that ought really, I suppose, to have been compared with kT above.

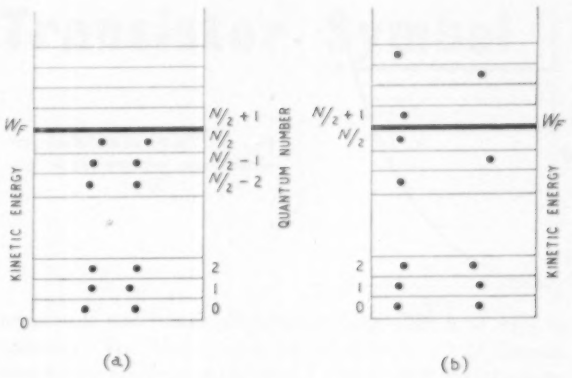


Fig. 2. *(a) At absolute zero. All the levels below W_F are filled with the complement of two electrons; no level above W_F is occupied at all; (b) above absolute zero. Some electrons have acquired sufficient energy to take them above W_F. The diagram exaggerates this effect, which only happens to a relatively small proportion of the electrons at ordinary temperatures*

Fermi Distribution at Finite Temperatures

The essential difference between the Maxwell-Boltzmann calculation and the Fermi-Dirac calculation is that, on the latter, the Exclusion Principle prevents more than two electrons entering the same cell in momentum space. What happens if there is so much energy in the system, and the momentum-space volume is so extensive, that there would be no tendency for more than two electrons to want to do so anyway, even if there were no Exclusion Principle? The answer is that, in this case, there is no distinction between the two kinds of statistics; and the distribution law for the electron gas at finite temperatures must be one that merges happily into the Maxwell-Boltzmann one if the temperature is high enough. The criterion for 'high' is that of the Boltzmann exponential term, $e^{-W/kT}$. As long as $W_F \gg kT$, which it is for any ordinary temperatures, this disorganizing term which, with its implication of lurking chaos does occur within the Fermi-Dirac formula, is relatively unimportant, and all that happens is a slight

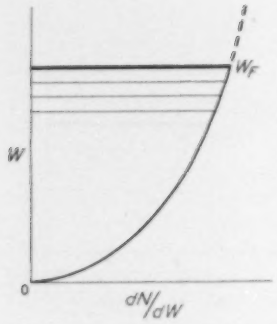


Fig. 3. *The Fermi distribution at absolute zero, corresponding to Fig. 2 (a)*

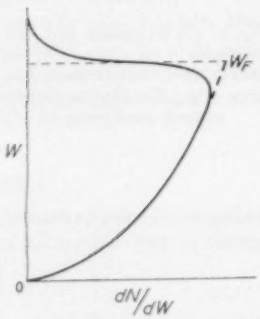


Fig. 4. *The Fermi distribution at temperatures above absolute zero, corresponding to Fig. 2 (b). By turning it through a right angle you can see a resemblance to the usual humped Maxwell-Boltzmann curve beginning; but, as with Fig. 2 (b), the rounding-off at ordinary temperatures is only slight*

rounding of the edges of the absolute-zero curve, which means that some levels above W_F are occupied (Fig. 4). If, however, kT is so large that it is of the same order as W_F, then chaos wins. But, as we have seen, this could only happen for T well above the melting-point of any metal; and, in fact, the actual distribution at ordinary temperatures differs little from that at absolute zero. This is one of the chief reasons why the Fermi Level is such a valuable piece of information; all sorts of assumptions remote from practice have had to be made in order to derive it; but it doesn't make very much difference if we discard them when they have served their turn.

Some Practical Points

The slight rounding of the distribution curve of Fig. 3 shows that there will be some abstraction of energy from the heat source by the electron-gas as its temperature is raised, and that there will therefore be some contribution by the electrons to the specific heat of the metal. This is in fact so; it is quite small, but measurements of the variation of specific heat with temperature agree with the calculated values.

You may ask, is there not something odd about a calculation that purports to deal with electrons, when no specifically electronic property (except m) has been called for?

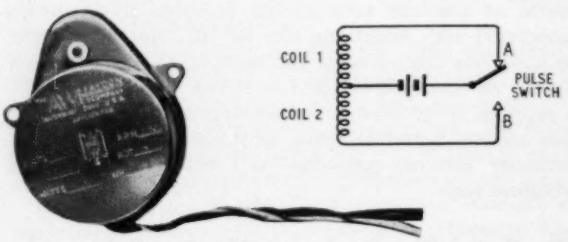
There are three answers to this. In the first place, electrons are the right kind of particle, with spin $\pm\frac{1}{2}$, for the statistics to apply to. In the second place, they are the only ones with the right-sized m (which appears in the denominator) to give a large W_F. And thirdly, they are the only kind of matter that we can really visualize as being gaseous at the unattainable absolute zero. We can, however, see what happens with the helium-3 isotope. This, comprising two protons and one neutron (each with spin $\frac{1}{2}$), does obey Fermi-Dirac statistics; the value of the mass to be put in the formula for W_F is roughly $6{,}000\,m$; the gas boils at about 3·2 °K, and this is as low as we can get really at ordinary pressure—when N/V is found from the density to be $2{\cdot}3 \times 10^{21}$ per cm^3. The appropriate W_F is then about 10^{-4} eV, corresponding to a little over 1 °K in kT terms; that is, W_F is *small* compared with kT for the actual temperature and, although the Fermi-Dirac statistics hold, we are already (although at the lowest physically reasonable temperatures) above the temperature at which they merge satisfactorily into Maxwell-Boltzmann statistics. In the case of helium-3, the proper calculation has its value, for it enables the rather large zero-point energy to be found; but it does not shed much light on the behaviour of this helium at ordinary temperatures. The reason why the electron-gas is so important a case is that the results *are* still relevant at ordinary temperatures; and this will be used when we start applying them to real electrons (which actually have charges) and to genuine metals (which are allowed to have an ionic lattice).

SMALL RATCHET-FREE STEPPER MOTORS

A small stepper motor which has been designed with two energizing coils to eliminate mechanical ratchets and which operates at a low voltage has been announced by The A. W. Haydon Company, Waterbury, Connecticut, U.S.A. The motor, Series 1800, does not require mechanical one-way devices, internal electrical contacts or a commutator.

When connected to an energizing source, as shown in the circuit diagram, one coil is energized in the motor; the rotor will advance 15° when the switch is transferred to the next position, energizing the second coil. Returning the switch to the first position will again energize the first coil advancing the rotor another 15° and completing one pulse cycle. Both coils must be alternatively energized to obtain rotation; consecutive pulses to one winding only will not advance the rotor. Pulse requirements are 12·5-msec minimum duration at up to 80 pulses per second.

Rated at 27 V d.c., the device has a continuous duty cycle, a rotor step angle of 30° per cycle and will operate over the temperature range of 85° to −54 °C. Load is stated as 0·2 oz in. at 20°C. It is capable of withstanding a 100-*g* shock for 7 msec and an acceleration 10 *g*. The rotor is completely surrounded by the iron stator which effectively encloses all rotating magnetic components.



A New Look at The Transistor Symbol

By R. van Winkle*

When I awoke from my twenty years' sleep into a world in which the vacuum tube had been all but replaced by the transistor and kindred devices, the new world, for an electronics engineer, took a lot of getting used to. I won't go into all the changes that had come about in techniques and equipment—radar, computers, atomic energy, network theory, the rise of the mathematician and the physicist, and so on—but there is one little thing in all this, a mere detail, really, which has continued to puzzle me, and on which I should like to spend a few words.

The little thing of which I write—and really I must apologize, for I feel I may be like the latecomer at a party who brightly raises a topic which had been completely exhausted in conversation just before he came in—the little thing is the *symbol* which is used to represent the transistor.

However, there is one fact that overcomes my hesitancy and strengthens my resolve and that is that now nearly three years after a set of symbols has supposedly been adopted as a standard by the I.R.E., one still sees used in the journals here and there any of the symbols shown in Fig. 1. And I find whenever I ask someone to explain to me in more detail how a certain device works, take a p-n-p-n device for example, he always begins by drawing something quite different from the symbol he uses. He almost always begins with a picture of a *block* of material, with various zones marked p and n. Hence I can't help concluding that, in the field of symbols at least, transistor engineers' are developing into marked schizophrenics.

The p-n-p-n device is an instance of a transistor type of device where the standardized symbol begins to break down. The I.R.E. symbol, which already appears to me as simply a little collection of sticks, now with a twig added is supposed to represent in some way the action of the p-n-p-n device. A good symbol, I have heard it said, should in some way be a picture of what it represents. This ensures two things: that its meaning will spring to mind without excessive exercise of the memory, and that it will be fertile in the sense that new variations of the thing symbolized will suggest new symbols in an organically related, rather than an arbitrary way. The criteria do not seem to me to be met by the little bundle of sticks.

In a way, I see, it *is* a picture of the thing symbolized. When one looks at a semiconductor device with the naked eye, all one sees is a tiny bar or flake of stuff with tiny wires sticking on to it from various directions. But to draw this is to convey nothing of the way it works, or why it was made, or what it does, that is to say, its *meaning*. To follow such an approach to its logical conclusion an organic chemist, to symbolize the structure of a molecule, might just as well draw a picture of a microscope, or of a beaker with pipettes dripping into it.

Now it is easy to criticize when one comes along after the fact, especially when one is an old man, grumpy from being awakened out of a sound sleep, and it is dangerous, as witness the way an immigrant is jumped on with both feet when he dares to criticize the new country he has wandered into, but does it not seem, even to the old inhabitant, that something got off the rails? Where then and how did it happen?

When I asked an old inhabitant about the origin of the commonly used symbol I was told that it arose from

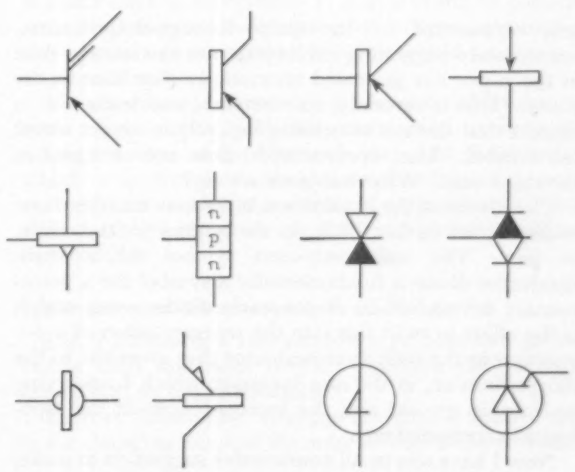


Fig. 1. *Some semiconductor-device symbols in current or recent use*

the first kind of transistor made, something called a 'point-contact' transistor, which is now apparently obsolete. Of this device, which consisted of two sharp wires lowered on to a slab of semiconductor, the symbol was a reasonable representation. But present types of transistor do not work quite the same way; in fact the point-contact transistor actually had a common-base current gain greater than one, a feature that would make many circuits as presently drawn unworkable—not just because of parameter differences but *in principle*. But, as I understand it, everyone continued to use the

* *Dr. R. W. Jackson, Director, Semiconductor Research Laboratory, RCA Victor Company Ltd., Montreal, Canada.*

original symbol either from force of habit, or from lack of ingenuity to think of a better one.

Some attempts were made to think of a better symbol, but all seem to have suffered the fate of nature's abortive mutations that lack adaptive advantage in the struggle for survival. One interesting example is shown in Fig. 2. It is the symbol proposed by P. M. Thompson and for a while adopted as standard by the Canadian Defence Research Board. It is derived from the diode symbol, in an obvious way, and it has the advantage that it suggests to the designer of switching circuits what is likely to happen when large voltages are applied one way and another to the device; i.e., it suggests that the collector junction is still a diode and can be forward biased. (The normally forward-biased junction, the emitter, is distinguished by shading.) Shading is a disadvantage, for quick drawing, but to my mind the real disadvantage shows up when one considers the symbol for the n-p-n transistor. What has happened to the base? Examining it carefully, one must conclude that the base region is

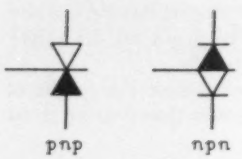


Fig. 2. *Canadian Defence Research Board transistor symbol*

now represented by the entire lozenge-shaped area, immediately suggesting intuitively that the storage time in the n-p-n has increased enormously over that in the p-n-p. This is certainly not the case, and leads one to suspect that there is something logically incorrect about the symbol. The representation does not *correspond* in the right way. What has gone wrong?

The source of the breakdown is obvious enough if one looks. It lies farther back, in the symbol for the diode, in fact. The commonly-used symbol for a semiconductor diode is fundamentally a symbol for a point-contact device, which all the early diodes were, and it is the effort to twist this into the representation of a p-n junction in the bulk semiconductor that gives rise to the illogicalities or, in the new language which I am trying to learn in my old age, the inconsistencies in the topological correspondence.

Now I have one small constructive suggestion to make, which I submit resolves all the difficulties. Represent the point-contact diode as before, as in Fig. 3 (a), but represent the junction diode as shown in Fig. 3 (b). It is obvious without explanation which is the forward direction of current flow, and which are the p and n regions. The symbol is *natural* enough.

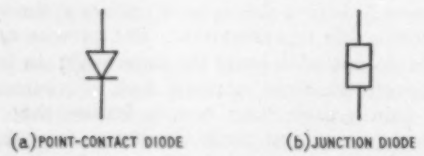


Fig. 3. *A suggested symbol for the junction diode*

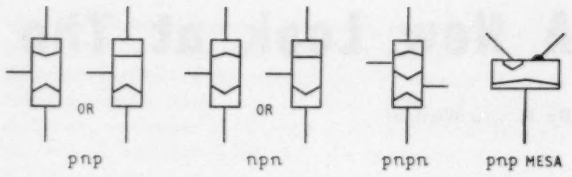


Fig. 4. *Examples derived from suggested junction symbol*

How to symbolize the junction transistor or, for that matter, any semiconductor junction device is now straightforward and obvious (Fig. 4), the true test of a good symbolism. I have added one refinement. The emitter junction can be distinguished by a deeper notch, suggesting a lower-impedance junction. In practice, for convenience, the collector junction can usually be drawn singly with a straight line. How one could draw a mesa type of construction is shown for good measure, in case there should be a reason for portraying more of the configuration of construction.

It would be foolish to expect the whole electronic community to leap at this suggestion and overthrow the ingrained habits and the masses of literature built up over the years. They are more likely to leap at the author, stuff his suggestion down his throat and cry "Back to bed, you old goat!" But it is only a suggestion, after all. I find that I can draw this symbol at least as quickly as any other, I do not confuse it with the wires and connections in the circuit diagram, and I do not have to draw *another* symbol when I want to explain how it works. And when it comes to symbolizing 'mol-ectronic' (ugh!) devices still more 'monolithic' than the p-n-p-n switch I shall be right in there building up symbols ad infinitum as I need them. I like my symbol. I hope you like yours, because if there is one thing worse than to spend 20 years asleep in bed, it is to spend 20 years asleep on one's feet.

EXPERIMENTAL MICRO-MINIATURE DIODE

An extremely small silicon diode, known as the Dot diode, has been developed by Hughes Aircraft in the U.S. and is to be produced in the U.K. by Hughes International.

The diode is hermetically sealed within a 0·050 in. diameter × 0·030 in. ceramic cylinder, closed at each end by a gold-clad metal plate. The end plates form the contacts for the diode and one of the plates is magnetized to facilitate handling the units.

At 25 °C, the breakdown voltage at 10 μA d.c., is 100 V d.c. min.; forward current at 1 V d.c., is 100 mA d.c. min.; and reverse current at 50 V d.c. is 100 mμA d.c. max.

The operating and storage-temperature range is 100 °C to −55 °C.

CORRECTION

The following corrections should be made to the article "Switching Transistors", by J. N. Barry, in the December 1960 issue:

In Fig. 12 (b), (c) and (d), p. 445, y_{ee} should be $y_{b'e}$.

P. 445, right-hand column, line 9, $r_{b'e'}$ should be $r_{b'e}$.

In the same column, $r_{b'e} = \frac{1}{g_{ee}}$ should be $r_{b'e} = \frac{\beta_o}{g_{ee}}$.

In the line below this, y_{ee} should be $y_{b'e}$.

Barkhausen Noise In Transformer Cores

By K. G. Warren, B.Sc.

The flux in a ferromagnetic core subjected to a changing magnetizing force varies partly by discrete steps and produces a noise component of the voltages in all windings linking the core. This is Barkhausen noise[1], and is attributed to the domain character of the material. A magnetic domain is a minute volume with a preferential axis of magnetization maintained by alignment of the electron spins among its atoms. A core consists of a multitude of these domains with their axes so disposed in the unmagnetized state that the external field is reduced to a minimum, this being achieved by the domains forming closed loops in the material.

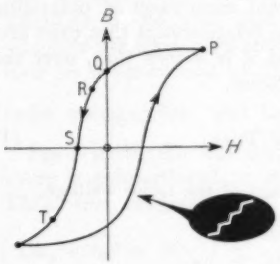


Fig. 1. *Illustrates that, in effect, the hysteresis loop of a ferromagnetic core is a staircase of small random steps*

The process of magnetization by an applied field has at least two distinct contributions. First, the boundary walls of a domain can move elastically within the confines of the inclusions (an inclusion is here considered as any inhomogenity in the material which will impede the motion of domain boundaries) and increase the overall magnetization. This action is reversible and free from noise. Secondly, the force on the boundary of a domain may be sufficient to cause it to jump beyond the restricting inclusion, partially embrace an adjacent domain and thereby increase the magnetization. This is a noisy process and irreversible in the sense that a subsequent reversal of the applied field will create a new domain pattern which is different, in detail, from the previous state.

In effect, a hysteresis loop is a staircase of random steps which alter each cycle but remain statistically unchanged. Fig. 1 illustrates this and the mechanism can explain some general features of the loop.

For instance, along PQ the field is reducing but has not reversed and there is no reversing force to alter the pattern of the domain boundaries, consequently the passage from P to Q is nearly noise free. On the other hand from R to S, where the field and the force have reversed, the action of domain break-up proceeds and from S to T they re-form in the opposite direction, and the passage from R to T is a noisy one.

It appears that an irreversible domain wall movement is very rapid[2] and probably limited by the effects of local eddy currents caused by the movement itself. However, the subsequent contribution to the main core flux is much slower since it is retarded by eddy currents throughout the core and by current flow in the windings linking the core.

Tebble[3] observed and explained a further effect which is of considerable significance in determining the spectrum of the noise. This is a minor avalanche that can arise. A small change in the applied field H may be just sufficient to bring about an irreversible wall movement which, in turn, increases the total field to an extent sufficient to cause other similar domain walls to move, and so on. This cluster of domain movements can greatly increase the grain of the noise and reference is made later to this effect. It is interesting to consider a material with an ideally square loop where, presumably, the switch between saturation levels would be due to a single avalanche.

We are concerned here with the isolation and measurement of the noise; with the relationship between the noise, the slope of the hysteresis loop and the speed at which it is cycled; with the introduction of a noise source in the equivalent circuit of a transformer; and with the frequency spectrum of the noise when polarized sinusoidally.

Isolation and Measurement of Noise

Fig. 2 illustrates a two-window core polarized on the centre limb and with similar windings, differentially connected, on the outer limbs.

Each of these noise windings develops a voltage due to the changing flux and the noise voltage is the deviation

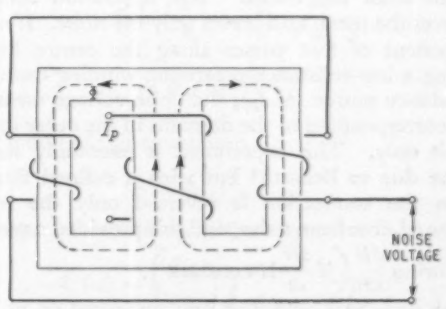


Fig. 2. *A two-window core polarized on the centre limb and with similar windings, differentially connected, on the outer limbs*

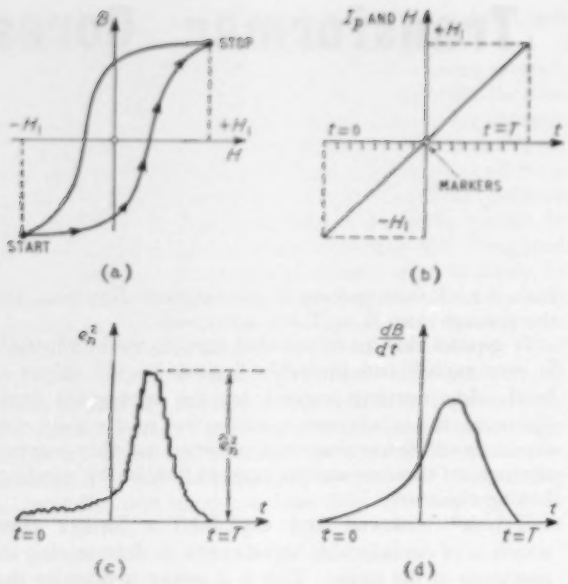


Fig. 3. *Outlines the ramp polarization experiment [$\hat{e}_n^2$ in (c) is the maximum value of the mean square noise]*

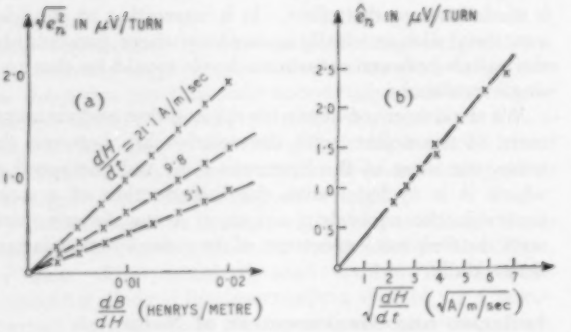


Fig. 4. *The relationship between the r.m.s. noise $\sqrt{e_n^2}$ and dB/dH (a) and at (b) the relationship between the maximum value of the r.m.s. noise $\hat{e}_n$ and dH/dt*

of this from the mean. The opposition connection removes the mean and leaves only the noise. If no noise component of flux passes along the centre limb, by feeding a low-resistance polarizing winding from a low-impedance source (of I_p), the noise voltage measured is that corresponding to the domains in the outer magnetic circuit only. This experiment is essentially the same as one due to Bozorth[4] but with a defined flux path. When the connection is reversed only the mean is measured, free from noise, and this provided a method of measuring $\frac{dB}{dH}\left(\text{if } \frac{dH}{dt} \text{ is constant}\right)$.

The noise was amplified and measured on an instrument designed to record the mean square value (e_n^2) with an averaging time constant of 0·2 sec, which was sufficient to give a consistent value yet short enough to follow the variations in the polarization. The overall bandwidth was increased until no further increase in output resulted so that all significant noise components were measured. The bandwidth extended from 40 c/s to 10 kc/s (3-dB points).

Ramp Polarization

The polarizing current I_p was supplied from a linear integrating unit with a constant-current input to produce a constant rate of current change from a negative reference, through zero, to an equal positive value. Time markers were included on the record and, before each sweep, the iron was demagnetized and carefully brought to the reference point. Fig. 3 illustrates the experiment.

Curves (c) and (d) were recorded photographically, and projected for analysis on to a graph screen. Since $\frac{dH}{dt}$ is constant, $\frac{dB}{dt}$ is convertible to μ, the permeability. Similar records were made for a series of time intervals T, and for different excursions of polarizing current I_p. Fig. 4 shows the relationships that exist and for the *sample of stalloy* used it is shown that, over the ranges involved, the r.m.s. noise,

$$\sqrt{e_n^2} = 20\sqrt{\frac{dH}{dt}}\,\frac{dB}{dH}\,\mu\text{V/Turn.} \qquad \ldots \quad \ldots \quad (1)$$

Fig. 5 shows an oscillogram of the noise voltage.

Deductions

$\sqrt{e_n^2}$ is an open-circuit r.m.s. noise voltage. It is proportional to the value of the permeability over a large area of the B-H plane investigated (no tests of small excursions about the origin were made). Furthermore, finding the r.m.s. noise voltage is proportional to $\frac{dB}{dH}$ rather than $\sqrt{\frac{dB}{dH}}$ would suggest that the permeability is controlled by the size of the discontinuities and not the density. Hence in the region RS in Fig. 1 the irreversible wall movements are larger than those in the region QR.

e_n^2 is proportional to $\frac{dH}{dt}$ and this was expected, since increasing $\frac{dH}{dt}$ is, in effect, climbing the Barkhausen steps more rapidly and causing more noise pulses in a given time interval. It follows that for sinusoidal polarization, in this range of speeds, the noise voltage is proportional to the square root of the frequency, whereas the mean voltage (i.e., the signal or polarizing voltage) increases directly as the frequency. Thus a signal-to-noise ratio is proportional to the square root of the signal frequency. It is emphasized that this does not hold at higher frequencies as will be seen later in this paper.

The constant in Equ. (1) is appropriate only to the core used in this experiment. It is reasonable to require a constant for a core material so that different metals can be compared on a noise basis. Thus the cross sectional area (A) and the length (l) should be involved

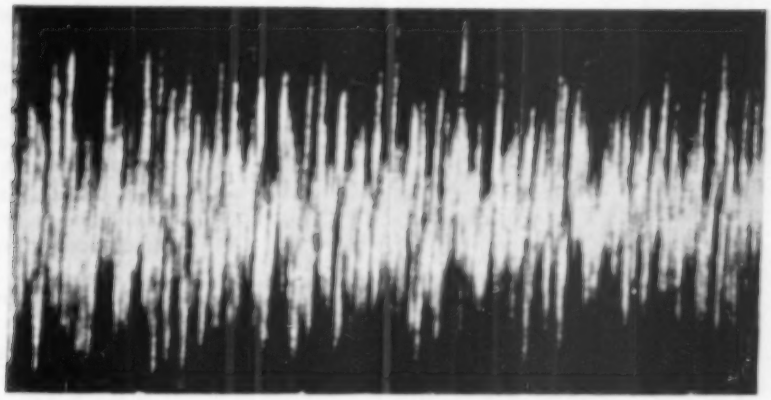


Fig. 5. *Oscillogram of the noise voltage*

and we might reason as follows. When a sample is taken from a flux density $-B$ to an equal value $+B$ a certain number of Barkhausen steps occur. If the volume density of these is ρ the average number is $lA\rho$. Each step contributes, on the average, an increment $2BA/lA\rho$ to the total flux. The average voltage step which is the time derivative of the noise flux will therefore be proportional to $2BA/lA\rho$, and the total r.m.s. noise voltage/cycle will be proportional to $\frac{2B}{l\rho}\sqrt{lA\rho}$.

For a particular flux density the r.m.s. noise voltage/cycle is proportional to $A^{\frac{1}{2}} . l^{-\frac{1}{2}}$ since ρ is then fixed. Thus the magnitude of the r.m.s. noise envelope $\left(\text{i.e., } \sqrt{e_n^2} = 20\sqrt{\frac{dH}{dt}\frac{dB}{dH}}\right)$ will also be proportional to $A^{\frac{1}{2}} . l^{-\frac{1}{2}}$. Dividing $\sqrt{e_n^2}$ by the value of $\sqrt{\frac{A}{l}}$ used in the experiment yields

$$\sqrt{e_n^2} = 676\sqrt{\frac{dH}{dt}\frac{dB}{dH}}\ \mu\text{V/Turn/}\sqrt{\text{m}}.$$

It must be noted that for a particular value of flux density the value of $\frac{dB}{dH}$ is fixed and known. The constant in the above expression is offered as the value for stalloy (0·015-in. laminations) providing the value of $\frac{dH}{dt}$ is kept low.

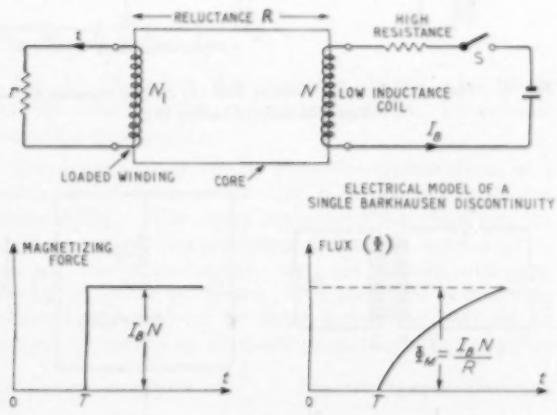


Fig. 6. *Electrical model of a single Barkhausen discontinuity*

Noise Generator in a Transformer Equivalent Network

Before attempting to place a noise generator in the usual equivalent networks of a transformer, it is necessary to find the effect of loaded windings on the Barkhausen noise. In order to carry out this analysis it is convenient to simulate the discontinuous boundary-wall movements by an electrical model. When a discontinuous domain-wall movement occurs it does so in a very short time and is equivalent to a current from a very high-resistance source being suddenly switched through a low-inductance coil wound round the specimen, the subsequent rise of flux in the core being controlled by eddy-current effects.

Let us consider one discontinuous boundary movement taking place in a core which is loaded with a single winding terminated in a resistance r.

The flux in the core at any instant after S is operated is

$$\Phi = \frac{I_B N - iN_1}{R} \qquad \ldots\ \ldots\ \ldots\ \ldots \quad (2)$$

$$\text{Also} \quad ir = N_1\frac{d\Phi}{dt} \qquad \ldots\ \ldots\ \ldots\ \ldots \quad (3)$$

From Equs (2) and (3)

$$\Phi = \frac{I_B N}{R} - \frac{N_1^2}{R}\frac{1}{r}\frac{d\Phi}{dt}$$

$$= \Phi_M - \frac{L_1}{r}\frac{d\Phi}{dt}$$

The solution of this equation is

$$\Phi = \Phi_M\left(1 - e^{\frac{-r}{L_1}t}\right)$$

The frequency transform of this expression is

$$\Phi(\omega) = \left(\frac{\Phi_m}{j\omega}\right) \times \frac{1}{\left(1 + j\frac{\omega L_1}{r}\right)}$$

where $\Phi_M/j\omega$ is the spectrum of the flux step if the

winding N_1 were on open circuit. The voltage across N_1 is

$$e = N_1 j\omega \times \Phi(\omega) = \frac{\Phi_M N_1}{\left(1 + j\frac{\omega L_1}{r}\right)}$$

$$\therefore |e| = \frac{\Phi_M N_1}{\sqrt{1 + \frac{\omega^2 {L_1}^2}{r^2}}} = \frac{e \text{ open circuit}}{\sqrt{1 + \frac{\omega^2 {L_1}^2}{r^2}}}$$

The above analysis shows that the most convenient form for the equivalent circuit is obtained by connecting a constant-voltage generator with a voltage equal to the open-circuit noise voltage in series with the shunt inductance of the loaded winding.

In order to show experimentally that the source impedance of the Barkhausen noise is the shunt inductance of the winding concerned two experiments were devised using the assembly shown in Fig. 2. In both, the core was polarized sinusoidally at 1 c/s and the noise output scanned by a Marconi Instruments wave analyser adapted to read r.m.s. with an averaging time of about 30 sec. In the first, the noise source was closed in a resistance R_L and the terminal voltage recorded as R_L was varied. The change in the spectrum due to R_L was also observed. Fig. 8 illustrates this experiment.

This curve and the observed tilt in the spectrum—the higher frequencies were de-emphasized by R_L—indicated that the source impedance was inductive; and of the order of that of the combined winding inductance. The difficulty is that the inductance varies with the permeability over the cycle of polarization. This introduced a degree of uncertainty and the second experiment closed one transformer in a second similar one, also polarized from the same source, and therefore the two inductances varied together. The load on each transformer was then the inductance of the other and, if the noise source impedance is due to this inductance value, the noise voltage from each should drop by one half and its spectrum remain unchanged. But the two are measured together at the same terminals and so the indicated voltage should be $(\frac{1}{2} \times \sqrt{2})$ of the open circuit noise e.m.f. This was confirmed and it is maintained that this shows the effective noise-source impedance to be that of the winding inductance.

To illustrate the application of this to a two-winding transformer let us consider the arrangement of Fig. 9 (a).

The polarizing current $|I_p| \simeq e \dfrac{\dfrac{R_L}{N^2}\Big/\left(\dfrac{R_L}{N^2} + R_S\right)}{\omega_p L_p}$ where L_p is the primary inductance, with N_p turns. The appropriate open-circuit noise voltage per turn $\sqrt{\overline{e_n}^2}$ is calculated or taken from spectrum measurements. Then with respect to noise only we can consider the network of Fig. 9 (b) where

$$V_n = \frac{\sqrt{\overline{e_n}^2} \cdot N_p \cdot R}{\sqrt{(R^2 + \omega^2 {L_p}^2)}}, \quad R = \frac{\dfrac{R_L}{N^2} \cdot R_S}{\left(\dfrac{R_L}{N^2} + R_S\right)}$$

and $\omega/2\pi$ is the frequency of the required noise component.

The secondary voltage is $N \cdot V_n$.

It is noted that since, in most applications, ωL_p is designed at its least value to be much in excess of R_S (or R_L), the noise voltage is necessarily small. This might account for the fact that many workers have not observed it at all, but it is stressed that $\sqrt{\overline{e_n}^2}$, the available noise e.m.f., is surprisingly large, as shown by the scales of Fig. 4.

Sinusoidal Polarization at Higher Frequencies

It was seen earlier that at low frequencies the noise voltage is proportional to the square root of the polarizing frequency and for stalloy this holds below about $\frac{1}{2}$ c/s. At higher frequencies the index is much reduced from one half. Experiments were made which measured the total (i.e., wide-band) noise voltage for a series of polarizing frequencies. The amplitude of the sinusoidal

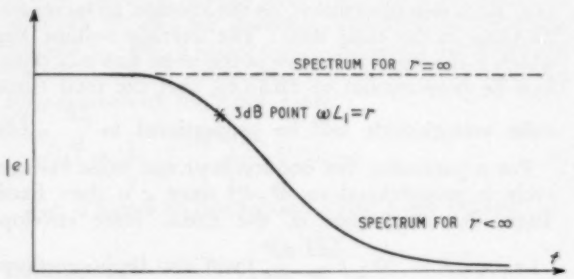


Fig. 7. *Shows how the spectrum is altered by loading the noise source with a resistance*

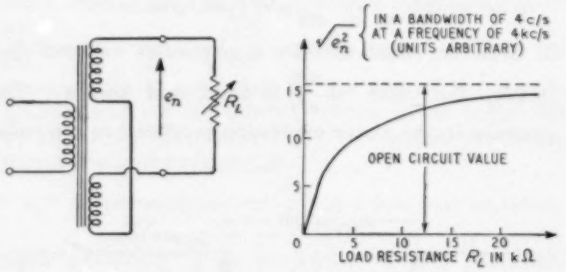


Fig. 8. Noise transformer on resistive load (a) and the variation of noise voltage with resistive loading (b)

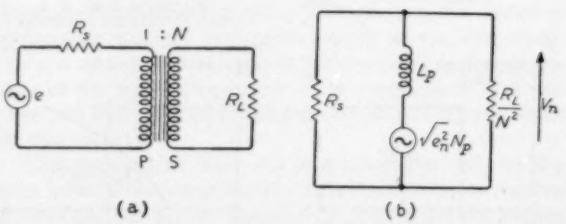


Fig. 9. *Typical transformer circuit (a), and the equivalent circuit for computing the magnitude of the Barkhausen noise (b)*

polarizing current was maintained constant throughout the experiment. The filtering problem is more difficult at higher frequencies due to greater effects from any unbalance of the two noise windings and the technique, in each measurement, was to widen the pass-band of the noise-measuring channel until no increase was indicated at the output, and then to disturb the balance and observe whether the output was affected. When it was not, we considered the noise isolation adequate. Fig. 10 shows the results of these tests.

From Fig. 10 it can be seen that e_n^2 is initially proportional to f_p up to $\frac{1}{2}$ c/s. Above this frequency the curve lies below the continuation of the original e_n^2–f_p curve. Further tentative tests at 200 c/s and 1 000 c/s suggest that at these higher frequencies the law of the curve is $e_n^2 \propto f_p$ but that the constant of proportionality in this range is very much less than in the range X–Y. The region Y–Z is therefore probably a region of transition from one constant of proportionality to another. If the character of the noise is the same at all frequencies the curve should show that $e_n^2 \propto f_p$ over all frequencies. The fact that it does not suggests that the noise mechanism undergoes a transition with frequency. This could be explained with reference to the 'avalanche action' previously mentioned. At low frequencies the avalanche action occurs when one discontinuity sets up sufficient force to cause another and so on, thus producing a relatively large irregularly-shaped pulse; the polarizing force increasing but little between the start and finish of the pulse. The noise waveform thus consists of a series of relatively-coarse pulses each made up of a very large number of discontinuities. As the polarizing frequency is increased up to $\frac{1}{2}$ c/s so the rate of occurrence of the pulses is increased and e_n^2 is proportional to f_p. Above 1 c/s it is suggested that the polarizing force is changing fast enough to initiate some of the discontinuities before the avalanche effect occurs. Thus, at higher frequencies, the avalanche effect is eliminated and the discontinuities are initiated mainly by the action of the polarizing force. The coarse avalanche pulses are replaced by a large number of much finer pulses which give rise to a much smaller e_n^2 than would have been so if the coarse pulses had prevailed at the higher frequencies.

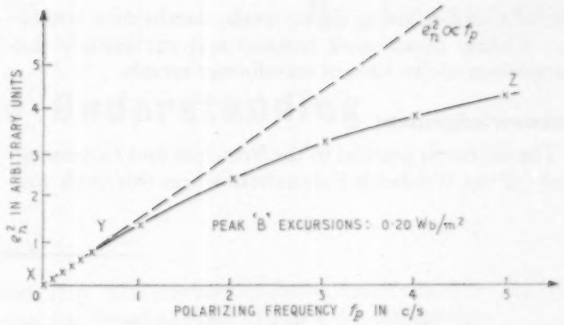


Fig. 10. *The relationship between the frequency of polarization and the mean square fluctuation of the noise*

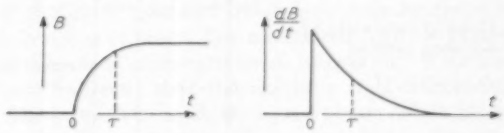


Fig. 11. *Flux and voltage pulses (τ is the time constant of the flux and voltage pulses)*

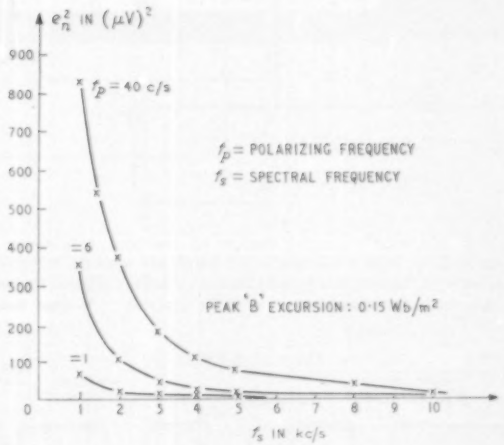


Fig. 12. *Frequency spectra of Barkhausen noise for three polarizing frequencies*

The Noise Spectrum

Fig. 11 illustrates the probable shapes of a single domain contribution subsequent to an irreversible boundary movement.

The noise voltage results from the superposition of a multitude of such pulses and is proportional to the permeability. The noise waveform has therefore an envelope whose instantaneous amplitude is characteristic of the permeability and the pulses, although sequential, occur randomly. The spectrum of the noise is largely determined by these factors, and would be continuous and decay inversely proportional to frequency for values in excess of $\frac{1}{\tau}$. Spectrum measurements were made for sinusoidal polarization using the wave analyser (4-c/s bandwidth) with a mean square indicating instrument and a long averaging time (30 sec.) The spectra are shown in Fig. 12. No line spectra were detected.

Conclusion

While it is not suggested that Barkhausen noise is of importance in many applications, there are those where it is a limiting factor, and it is clear that transformers can be designed and operated so as to introduce only a small noise contribution. Furthermore, the noise becomes calculable if the circumstances under which the transformer will be used are known and this itself

can be useful in fixing signal levels, bandwidths and so on. Clearly much work remains and particularly the comparison of the various transformer metals.

Acknowledgement

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REFERENCES

[1] H. Barkhausen, "Two Phenomena Uncovered with the Help of the New Amplifiers", *Z. Phys.*, 1919, Vol. 20, p. 401.
[2] R. Becker, *Z. Tech. Phys.*, 1938, Vol. 19, p. 542.
[3] R. S. Tebble, "Barkhausen Effect in Single Crystals", *Proc. phys. Soc.*, 1953, Vol. 68, p. 633.
[4] R. M. Bozorth, *Phys. Rev.*, 1929, Vol. 34, p. 772.

Turret Tuner For V.H.F. And U.H.F.

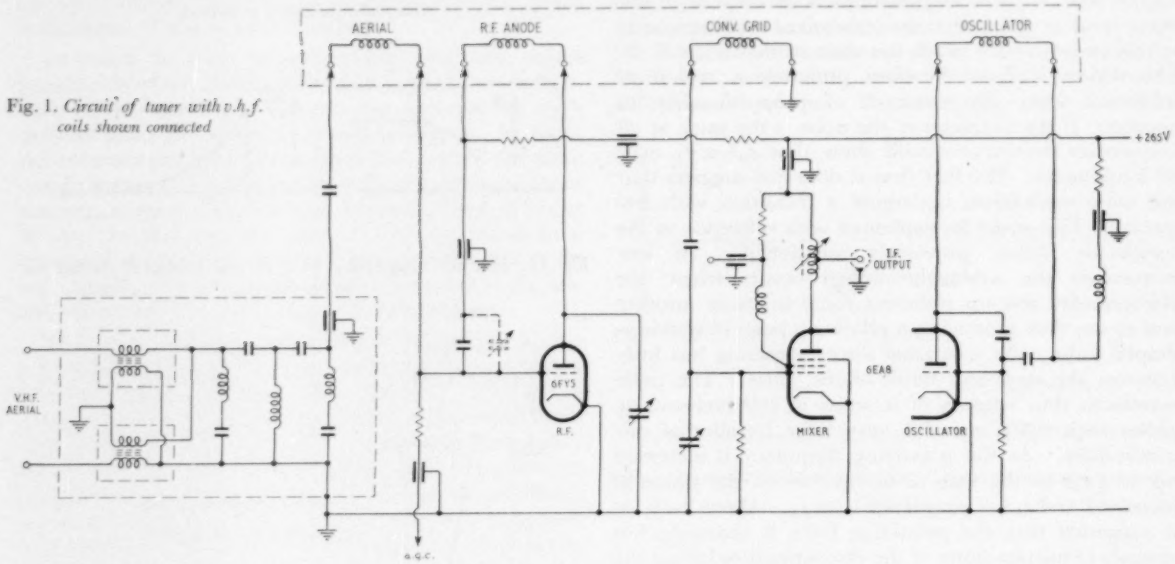


Fig. 1. *Circuit of tuner with v.h.f. coils shown connected*

In the U.S.A., both v.h.f. and u.h.f. bands are used for television and, in view of the suggestion that Bands IV and V should be used here, American practice is not without interest. A new tuner announced by the Zenith Radio Corporation can carry u.h.f. 'biscuits' as well as v.h.f. in the tuner. On v.h.f., there is a neutralized triode r.f. stage with a 6FY5 valve and a triode-pentode frequency-changer. Coupled circuits are used between the valves and the aerial input (300-Ω balanced) is taken through a balun and low-pass filter. Otherwise, the circuitry is fairly conventional and takes the form shown in Fig. 1.

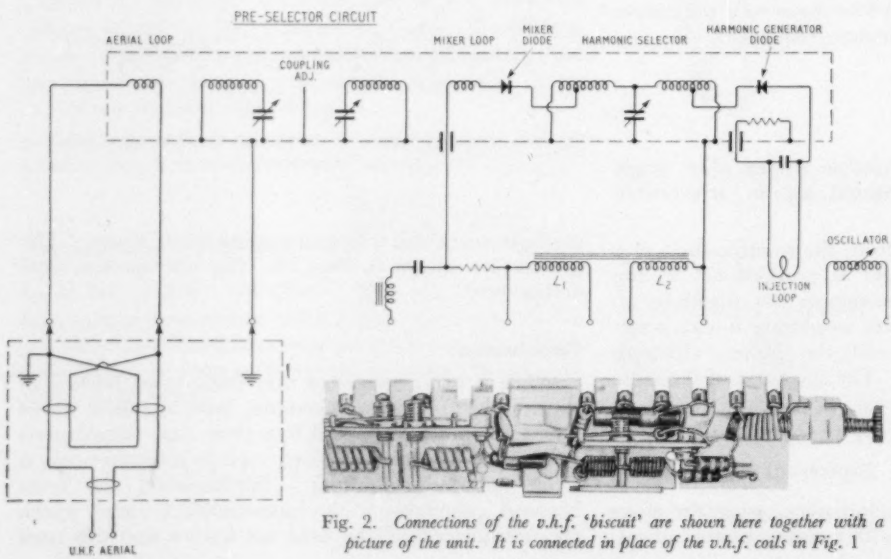


Fig. 2. *Connections of the v.h.f. 'biscuit' are shown here together with a picture of the unit. It is connected in place of the v.h.f. coils in Fig.* 1

On u.h.f., the arrangement shown in Fig. 2 is brought into play. Three extra contacts are needed on the 'biscuit' which occupies twice the space of a v.h.f. 'biscuit'. The oscillator is coupled into a crystal diode which acts as a harmonic generator and a tuned circuit selects the required harmonic and feeds it to the mixing crystal diode to which the signal is applied via a coupled pair of tuned circuits. The i.f. signal produced is fed to the grid of the triode and the coils L_1 and L_2 form the intervalve coupling to the pentode, these valves acting on u.h.f. as extra i.f. stages. A sketch of the u.h.f. 'biscuit' appears in Fig. 2.

Routes to Creative Understanding

For some time now in this series of articles we have been advocating the importance of operational calculus in which the symbol p is identical with d/dt and is not a mere number, although it can be handled as if it were a number; for example, when a rational function (quotient of two polynomials) in p is split up into partial fractions. We have thus advocated operational calculus partly because it is more general than possible alternative techniques such as Laplace transforms, and partly because of an unfortunate historical accident. Heaviside, who knew intuitively himself exactly what he could do with his operator p and what he could not do, could not be bothered to explain his methods and ideas to the satisfaction of contemporary Cambridge mathematicians. Heaviside was concerned with achieving understanding of actual phenomena, and his very success in doing this by what the Cambridge mathematicians regarded as a series of fouls deserving the extreme wrath of any self-respecting referee caused these mathematicians to become his enemies. Once they had become Heaviside's enemies, they devoted their energies to proving that Heaviside was unsound—a purely negative and destructive objective. It is unfortunately a natural human reaction to many useful new ideas to seek to prove they are wrong (because it is such a bother to change existing ideas to give the new idea a fair hearing). Unfortunately, this natural reaction contains significant elements of pride, that deadliest of the seven deadly sins. Here, however, we are only concerned with trying to deal with the consequences of this natural reaction. These consequences are that the engineer of 1961 is presented with a variety of possible techniques for solving his problems (which means reaching creative understanding about them). These techniques include Laplace transforms, symbolic calculus, the impedance concept, matrices, Fourier transforms, etc., and this article is devoted to discussing in general terms the way in which all these various techniques fit together to help in the ultimate objective of reaching creative understanding. The position might be summarized by Fig. 1, of which the rest of this article is an explanation.

In any practical problem, we are given a 'machine or situation' which is real and existent in the actual world, and our ultimate objective is to reach creative understanding thereof. Creative understanding does not necessarily mean that we have to be given the input as a function of time and are obliged to derive the output as a function of time. The machine might, for example, be an idealized differentiator or integrator. If we know, or can find out, that the machine is a differentiator, we know all we need to know about what it does. Given any particular input, we can derive the corres-

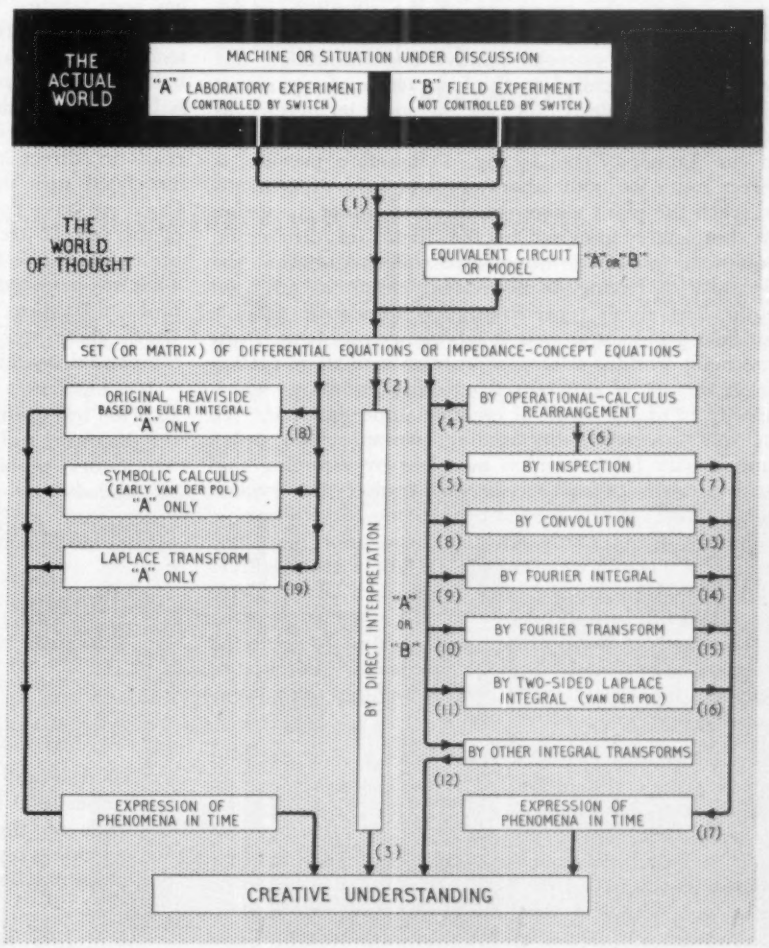


Fig. 1. *Routes to Creative Understanding*

ponding output by a process trivially simple compared with this general understanding that the machine will always differentiate the input and deliver the derivative as output. This kind of direct route to creative understanding is shown at the point numbered (2) in Fig. 1, but some preliminary explanation is required.

Given any machine or situation in the real, actual world, our objective is an expedition into the world of thought to achieve creative understanding (depicted at the bottom of Fig. 1). The expression 'laboratory experiment' in Fig. 1 is used to denote any kind of experiment, observation or investigation in which the whole machine is initially dead and is galvanized into action suddenly at a time, say t_0, under the control of the observer, perhaps by the closing of a switch. The expression 'field experiment' is used to denote experiments or investigations where the observer has no control over the phenomena investigated, as in astronomy for example. The junction points where a choice of route is possible are all numbered in Fig. 1 for easy reference only. For either type of experiment, we can cross the boundary into the world of thought and proceed to formulate an equivalent circuit [see (1) in Fig. 1] and thus derive a set of differential equations which specify the properties of the situation under discussion completely, but in a language which may not immediately provoke creative understanding. But we are not compelled to think in terms of equivalent circuits and, indeed, the use of such techniques as matrices may make it unnecessary to do so, as indicated in the January 'Tool'. Fig. 1 therefore provides a direct route for both laboratory and field experiments to reach the point where the situation under discussion is completely specified in the idiom of differential equations (or of the impedance concept) without the necessity of having an equivalent circuit at all. An equivalent-circuit representation is sometimes possible for field problems as well as laboratory problems. Now the main work of solving a problem consists in getting from the set of differential or other equations which specify it to the stage of creative understanding. To achieve creative understanding, we must not inflict unnecessary blinkers upon ourselves by assuming that such understanding necessarily requires the explicit formulation of all relevant phenomena in terms of functions of time, although such formulation may often be useful. That is why the route DIRECT INTERPRETATION is provided in Fig. 1, in which all numbered junctions can be by-passed [after (2)] to achieve creative understanding, as in the case of the differentiator already mentioned, without necessarily formulating explicit time-functions to represent input or output at all.

Another case in which the 'direct interpretation' route is the best is illustrated by the example where the differential equation is $y = [p/(p^2 + ap + b)]\ H(t)$ where a and b are positive constants, $p = d/dt$ and $H(t)$ is Heaviside's unit step. Here the useful information is contained in the fact that a and b are positive; y as a function of time must necessarily contain two terms decaying exponentially, with oscillation if $a^2 < 4b$ and without oscillation if $a^2 > 4b$. This is often all we need to know, and the laborious determination of y in terms of sines and cosines often gives little, if any, useful additional information.

When we are not fortunate enough to be able to obtain 'direct interpretation' at the differential-equation stage, it is very often possible to obtain the required expression in terms of explicit time functions by inspection of the differential equations (or equivalent equations, or matrices of coefficients thereof, obtained by using the generalized impedance concept in which the impedance of a pure inductance is pL, etc.) if necessary, after preliminary manipulation (such as resolution into partial fractions). This is symbolized in Fig. 1 by the route to creative understanding via the junctions (4), (6), (7) and then vertically down to (17) and explicit time functions; the case in which the solution in time functions is obvious is symbolized by the route (4) (5) (7) and thence vertically through (17). We contend that along both these routes p is essentially an operator, equivalent to d/dt, and that any attempt to apply the relevant operator to an operand should be deferred until the last possible moment—in fact, until we are sure we have manipulated the operator into the most convenient form available.

That is why the block between junctions (4) and (6) has been labelled "BY OPERATIONAL-CALCULUS REARRANGEMENT". It is possible to use other techniques as indicated in lower routes on the right-hand half of the diagram, and in both routes on the left-hand half, but we maintain that the route (4) (5) (7) or (4) (6) (7) is the best available in most cases, because roads to 'direct interpretation' of some terms at least associated with the problem may thus be opened up. Once we have committed ourselves to the use of any of the specific techniques mentioned in other parts of the diagram, we lose the possibility of a 'short cut' by what we have called 'direct interpretation', unless the problem happens to belong to the restricted classes for which these specific techniques are necessary; e.g., differential equations involving fractional powers of p.

We have now to explain why we have relegated Laplace transforms and symbolic calculus to the left of the diagram, and regarded them as applicable only to 'laboratory' problems. This is effectively a point that has already arisen in an apparently different form. In order that algebraic manipulations of operational expressions (functions of p) should be valid, it is necessary that

$$p\,h(t) = \frac{d\,h(t)}{dt}\ ;\quad p^{-1}\,h(t) = \int_{-\infty}^{t} h(\tau)\,d\tau \qquad \ldots \quad (1)$$

so that

$$p\,p^{-1}\,h(t) = p^{-1}\,p\,h(t) = h(t) \qquad \ldots \quad \ldots \quad \ldots \quad (2)$$

for all relevant operands $h(t)$, as indicated in the September 1960 Tool. But, unfortunately, the definition -integrals of the Laplace transform $F(s)$ and the corresponding representation $f(p)$ in symbolic calculus of a given function of time $h(t)$ are respectively

$$F(s) = \int_0^\infty e^{-st}\,h(t)\,dt\ ;\quad f(p) = p\int_0^\infty e^{-pt}\,h(t)\,dt \qquad \ldots \quad (3)$$

and this has an effect which is equivalent, in the language

of operational calculus, to replacing the second of Equs. (1) by

$$p^{-1}\,h(t) = \int_0^t h(\tau)\,d\tau = \int_{-\infty}^{t} h(\tau)\,H(\tau)\,d\tau \quad .. \quad .. \quad (4)$$

where $H(x)$ denotes Heaviside's unit step, zero for negative x and unity for positive x. It follows that the operand is really $h(t)H(t)$ in Equ. (4), and thus (one-sided) Laplace-transform calculus and symbolic calculus are strictly only applicable to functions of time which contain a step-function as a factor; such functions of time are associated with what we have called 'laboratory experiments'. On the other hand, van der Pol and Bremmer have studied transforms based on what they call the two-sided Laplace transform (reference 1) and, in this case, the restriction of applicability to 'laboratory experiments' is no longer valid, so this possibility is included in the route of Fig. 1 via (11) and (16) and is legitimately on the right-hand side of that figure.

The route via (8) and (13), marked BY CONVOLUTION in Fig. 1, is a perfectly valid one and has the advantage that all operations leading to creative understanding are carried on in the time world, and no transforms are required. If $x(t)$ is the input to a system whose response to unit impulse is $\phi(t)$, then the output is $y(t)$ where

$$y(t) = \int_{-\infty}^{\infty} x(t)\,\phi(t-\tau)\,d\tau \quad .. \quad .. \quad .. \quad (5)$$

and the integral on the right-hand side of Equ. (5) is well known as the 'convolution integral' of the two constituent functions x and ϕ; these may be exchanged without affecting $y(t)$. The disadvantage of this 'convolution' route is that the operation inverse to Equ. (5), of finding say $x(t)$ when $y(t)$ and $\phi(t)$ are given, is not straightforward, whereas the corresponding process in operational calculus, or Laplace-transform or symbolic-calculus technique is greatly assisted by using all the resources of the theory of functions of a complex variable, applied to the operator or transform in question. There remains the question of the Fourier route, via (9) and (14) or (10) and (15) in Fig. 1. Here the main difficulty is the question of inadequate specification discussed in recent 'Tools': in the original form of the Fourier integral it is as if a limit (associated with the universal presence of stray elements) has been prematurely closed, and we have already warned readers repeatedly of the grievous consequences of errors of this nature. This warning applies mutatis mutandis to all methods and techniques; a question of essence, not of method, is involved.

Thus, while the operational-calculus approach is clearly not the only possible approach and, at present, by no means the most usual, we contend that it is the one to try first, because it is the most general (as we have tried to show in general terms) and has the greatest possibilities of a short cut to the only significant objective of creative understanding.

REFERENCE

[1] B. Van der Pol and H. Bremmer, "Operational Calculus based on the Two-Sided Laplace Integral", Cambridge University Press, 1950 and later editions.

CAMBRIDGE MICROANALYSERS FOR U.S.

The first of four Cambridge Instrument microanalysers ordered by organizations in the United States has recently been shipped to the Illinois Institute of Technology in Chicago. The following two are destined for the United States Steel Corporation and The Crucible Steel Company of America. It is stated that these instruments were selected by the various American organizations because of a new British scanning technique.*

The technique of X-ray microanalysis consists of focusing an electron beam, normally accelerated from 0 to 50 kV, on to a region of the specimen surface approximately 1μ in diameter and rather less in depth, and analysing spectroscopically the resulting X-rays excited from the specimen. So far, the technique has been used almost exclusively for research work on metallurgical problems, although it has applications in other fields.

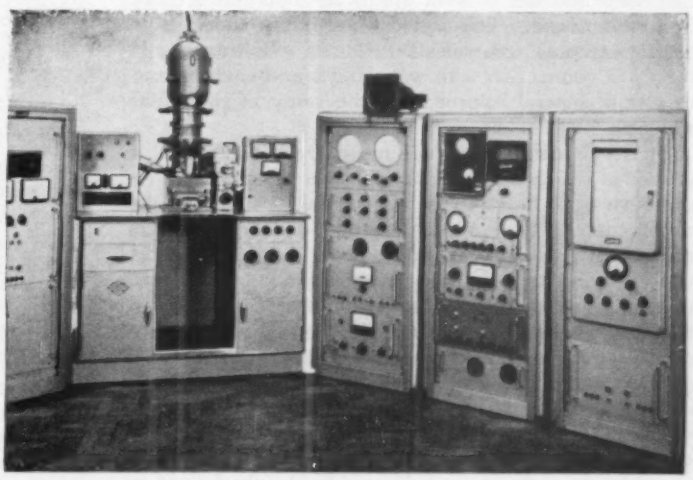
As well as static-probe point-to-point quantitative microanalyses, the Cambridge Instrument can also be used for scanning the surface of an area up to $\frac{1}{2}$ mm $\times$ $\frac{1}{2}$ mm. The characteristic X-rays from any single element (with atomic number 12, or greater) can be selected by a crystal spectrometer and detected by a proportional counter. The signal from the counter modulates the brightness of a beam scanning a c.r.t. in synchronism with the probe scanning the specimen and forms an image corresponding with the area scanned and showing semi-quantitatively the distribution of the chosen element.

Some of the back-scattered electrons from the surface of the specimen strike a scintillation counter, the signal from which modulates the beam of a second c.r.t. to form a reflected electron-image of the region. This region is similar in its quality of contrast to an optical image photographed by oblique illumination and shows topographical details.

The magnification for both images can be controlled from $\times$ 200 to $\times$ 3000, and the resolution is about 1μ although this can, under good conditions, be reduced to $\frac{1}{2}\mu$ for the electron image.

The accuracy for quantitative analysis is $\pm$ 1% which improves, under favourable conditions, to 0·1%. The minimum concentration that can be detected under good conditions is 0·1% corresponding to a mass sensitivity of about 10^{-14} gramme.

* The Cambridge Scanning Electron-Probe X-ray Microanalyser is based on one designed by Dr. D. A. Melford and colleagues at the Tube Investments Research Laboratories with the co-operation of Drs. V. E. Cosslett and P. Duncomb, who were responsible for the first model developed in the Cavendish Laboratory, Cambridge.



An illustration of the complete Cambridge microanalyser including the c.r.t. display units and the pen recorder

The Theory of Waveguides and Cavities

1: A GENERAL APPROACH TO THE EXACT THEORY OF WAVEGUIDES AND CAVITIES

By R. A. Waldron, M.A. (Cantab.), A.Inst. P., A.M. Brit. I.R.E.*

SUMMARY. *An outline is given of a general approach to the study of waveguides and cavities. The mathematics is not given in detail, but the difficulties encountered in setting up the mathematical problem for a given physical situation, and the physical principles involved in the treatment, are discussed at length. The method is to derive a wave equation from Maxwell's equations, and to solve it. Boundary conditions are applied to the solutions, and a number of equations are obtained; the expression for the consistency of these equations gives the characteristic equation. This can be solved, in principle, for the phase constant although, in practice, this is often very difficult.*

Many textbooks give the simple theory of waveguides and cavities, which is a part of the general education of all electrical engineers and many physicists and mathematicians. This theory deals with guides in the form of metal tubes and cavities in the form of metal-walled enclosures, both guides and cavities normally being filled with ordinary air. In this theory, certain features are observed which hold generally in the limited field to which the theory is applicable, namely the simple guides and cavities mentioned. The simplicity of these properties gives them a certain appeal, so that they become second nature to the microwave scientist.

However, in recent years more complicated waveguides and cavities have become of great importance, and it has been found that the features of simple guides and cavities which are so well known do not apply in the new circumstances. An understanding of the theory of more complicated guides and cavities usually has to be gained laboriously through long experience, and even so there is much evidence in the literature that some features of such guides and cavities are not fully understood. The aim of the present article is to discuss certain general features of electromagnetic theory in connection with waveguides and cavities, and to give a general approach to the study of particular systems.

In the following we shall mean by a waveguide a system in which all boundary surfaces are parallel to a given straight line, which will be taken to be in the z direction. The media bounded by these surfaces may be of any kind – metallic, dielectric, or magnetic, the permittivity and/or permeability being real, imaginary, or complex scalars or tensors. It is essential that there be at least one surface dividing at least two media if the system is to be a waveguide; in the absence of any such surface, we have free space or, what is equivalent, an infinite dielectric or other medium, in which plane waves can propagate in any direction. In this article we are concerned only with waves which travel in the z direction, being guided by the boundary or boundaries between various media. The boundary surfaces may be of any form, as long as they can be generated by a straight line which moves always parallel to the z axis. Cylinders of circular and rectangular cross-section immediately spring to mind; other shapes are possible, but are less frequently used. A waveguide may have more than one boundary surface, for example a dielectric-coated wire, or a conventional waveguide containing a ferrite or dielectric rod.

By a cavity we shall mean the portion of a waveguide, as defined above, bounded by the plane surfaces $z = z_1$, $z = z_2$, normal to the z axis. This is a restricted type of cavity, chosen for convenience in the development of the discussion. More complicated cavities, e.g., a spherical cavity, can be treated by the same general approach as for the cavities that will be discussed.

Preliminary Questions

The Choice of a Co-ordinate System

It is convenient to work in a co-ordinate system which is related to the symmetry of the system under consideration, if this is possible. The position of a point in space can be represented by three co-ordinates, ξ, η, z, and the study of a particular waveguide or cavity is simplified if its cylindrical surfaces are described by equations of the form $\xi =$ constant, $\eta =$ constant. For example, if the problem is to treat a metallic waveguide of circular cross-section containing a concentric rod of some material – dielectric, ferrite, or another metal – also of circular cross-section, the obvious co-ordinates to choose are the circularly cylindrical system, with $\xi = r$ and $\eta = \theta$. The two boundary surfaces are then given by $r = r_1$, $r = r_2$.

Idealization of Materials

A practical problem is often much simplified by idealizing the materials that are present. If a dielectric has a very small value of loss tangent, it can, to a first approximation, be replaced for the purposes of calcula-

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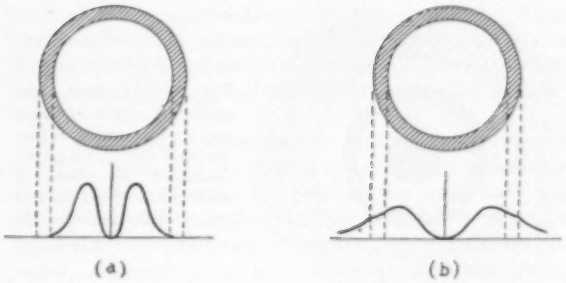

Fig. 1. *Power distributions in (a) metal-tube waveguide; (b) dielectric tube waveguide*

tion by a lossless dielectric of the same dielectric constant. A metal can be regarded as a perfect conductor. The effects of the small loss tangent of the dielectric, or of the large but finite conductivity of the metal, can be taken into account by perturbation methods, which are valid under practical conditions if the distortions of the field components that are produced by the approximations are small. Similar methods can be applied to ferrites and plasmas. A very lossy dielectric, however, cannot be idealized in this way; it is too lossy to be regarded as an approximation to a lossless dielectric, and not a good enough conductor to be taken as an approximation to a perfect conductor. An example of such a material is a plastic dielectric such as polythene, with powdered graphite embedded in it.

The Choice of a Model

No problem that arises in nature is ever susceptible of an exact mathematical treatment. An actual problem has to be replaced for purposes of calculation by an idealized problem which cannot arise in practice. The art of the theoretical physicist is to choose a model that is amenable to mathematical treatment – as general as possible, and preferably exact – and, at the same time, is sufficiently similar to the actual physical system to give results that are relevant.

The first point to be decided is the co-ordinate system in which the problem is to be treated. In practice, the surfaces of the media are never perfect geometrical surfaces, but they are sufficiently nearly so for a treatment in which they are taken to be geometrical surfaces to be valid. Next, the outermost medium is taken as extending to infinity; this is a safe assumption if the fields are negligible at the actual outer limits of this medium. Thus for a metal tube, very much greater than the skin depth in thickness, the assumption that it is infinitely thick is valid because the electromagnetic fields will be mainly confined within the tube, and only a very little energy will penetrate a very little way into the metal (Fig. 1). But if the tube is of dielectric, a considerable proportion of the electromagnetic energy will travel outside the tube, and it is no longer justified to take the outer surface to be at infinity. Rather, the air surrounding the dielectric tube must also be regarded as part of the system; if there are no obstacles sufficiently near to the tube for the fields to be appreciable, the air may then be regarded as extending to infinity.

It is often convenient to consider the media to be lossless, although this may not be necessary. Consider, for example, a simple waveguide consisting of a metal tube containing air. In the elementary theory, the metal is taken to be a perfect conductor, although the problem does not become excessively difficult if the finite conductivity of the metal is taken into account. Air is almost always taken as free space; not only are its properties sufficiently close to those of free space for all practical purposes, but the difference between its properties and those of free space will vary considerably from one sample of air to another.

The aim in choosing a model is always to use a model whose mode spectrum, it is reasonable to suppose, does not depart greatly from that of the system actually existing. Each mode of a waveguide has a characteristic eigenvalue, the phase constant β, which is 2π divided by the guide wavelength, and a characteristic set of field components. It is such a set of field components, with its eigenvalue, that we mean when we speak of a mode. Clearly, the field patterns and guide wavelength of a metal tube containing air are not greatly different from those of a hole containing vacuum in a perfect conductor extending to infinity. The latter is therefore a suitable model for the mathematical treatment of the former.

This model may still be taken if the tube contains also a thin rod of some material at its centre, provided that the rod is very thin or that its dielectric constant and permeability do not depart greatly from those of free space. The presence of the rod will distort the field patterns, but they will still be sufficiently similar to those of the model for the various modes to be recognizable, and the distortion can be calculated by perturbation methods. If the rod is thick, however – say, more than a fifth of the guide radius – and the dielectric constant considerably different from unity, the distortion can be so great that it is not possible to see the similarity between the field patterns of corresponding modes, and the value of the phase constant will not be accurately given by perturbation theory (Fig. 2). In this case it is necessary to choose a new model, and the most obvious one is a perfect conductor extending to infinity, with a a hole containing a rod of, e.g., lossless dielectric or ferrite, the space between the rod and the perfect conductor being occupied by vacuum. The mode spectrum for this

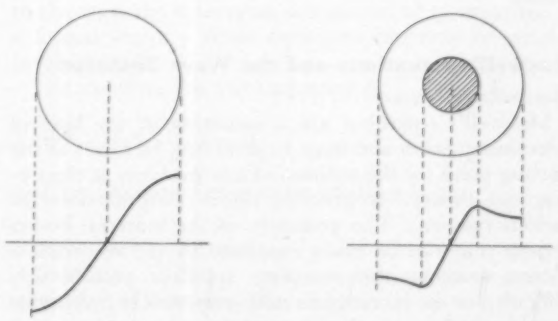

Fig. 2. *Longitudinal magnetic-field component in waveguide with and without concentric dielectric rod*

system has been examined in detail[1], and it has been shown that to each mode of the structure without the rod corresponds a single mode of the structure with the rod, and vice versa.

Nomenclature

Before proceeding further, it will be convenient to define certain types of waveguide. These definitions are not in general use, but there is a need for such definitions, and they will be used for the purposes of the series of articles of which this is the first.

By a simple waveguide will be meant one in which only scalar media are present, such as metals, dielectrics, vacuum, or air, but not ferrites or plasmas (Fig. 3). The media may be lossless or not. We shall describe as perfect a guide in which all the media are lossless. Thus a perfect guide can contain ideal lossless dielectrics or ferrites, or perfect conductors. It is an idealization not realizable in practice, although it can be approached very closely. Finally, a guide will be described as homogeneous if the energy of the electromagnetic wave is confined to a single homogeneous medium (Fig. 4). To confine the energy to one medium, the medium outside the boundary surface must be a perfect conductor. We shall denote the possession by a waveguide of the properties of homogeneity, simplicity, and perfection, in the above senses, by the letters H, S, and P, respectively. Clearly, a guide may have none, any one, any two, or all three of the properties H, S, P, in theory if not in fact.

The most elementary kind of waveguide is one consisting of a single boundary surface, outside which is a perfect conductor and inside which is a single, lossless, isotropic, dielectric material. This is the HSP guide, which is commonly used as a model for the simplest practical guide, the metal tube containing an isotropic, homogeneous, slightly lossy dielectric medium. The theory of the HSP guide is well known, and many concepts have arisen in the development of this theory, e.g., the idea of E and H waves, impedance, and the Poynting vector. These concepts colour the outlook of many microwave scientists, which makes the examination of non-HSP guides more difficult than it need be. We shall see in the second article of this series, where certain examples will be treated, that our ideas about waveguides, formed in the context of the HSP guide, require revision when considering more complicated guides.

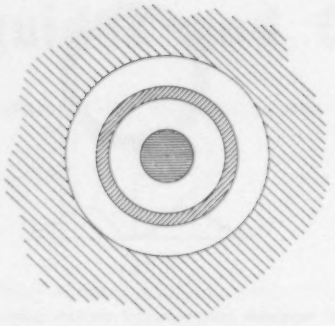

Fig. 3. *In a simple waveguide, any number of media may be present as long as they are all scalar. In a perfect waveguide, any number of media may be present and they may be scalar or tensor, but they are all lossless*

Maxwell's Equations and the Wave Equation

Maxwell's Equations

Maxwell's equations are a summary of the laws of electromagnetism and may, in principle, be taken as the starting point for the solution of any problem in electromagnetic theory. In practice, this is often not done for various reasons. The geometry of the material bodies present may not be easily expressed in any co-ordinate system, making the complete solution prohibitively difficult; or the investigator may only wish to investigate a single mode, e.g., the transmission-line mode, which can be treated by network theory. However, whenever it is possible, and the time required is not excessive, it is desirable to use the approach from Maxwell's equations since in this way a complete set of solutions can be obtained. Even though only one mode may be used in practice, it is often helpful to know how the other modes behave, if only for the purpose of avoiding them.

Maxwell's equations are, in their most general form,

$$\text{Div } \mathbf{D} = \rho \quad .. \quad .. \quad .. \quad .. \quad .. \quad (1)$$

$$\text{Div } \mathbf{B} = 0 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (2)$$

$$\text{Curl } \mathbf{H} = \sigma\mathbf{E} + \partial\mathbf{D}/\partial t \quad .. \quad .. \quad .. \quad (3)$$

$$\text{Curl } \mathbf{E} = -\partial\mathbf{B}/\partial t \quad .. \quad .. \quad .. \quad .. \quad (4)$$

Here ρ is the charge density at the point where the divergence is taken, and σ is the conductivity of the medium. We have further the relations, for media other than free space,

$$\mathbf{D} = \epsilon\epsilon_0\mathbf{E} \quad .. \quad .. \quad .. \quad .. \quad (5)$$

$$\mathbf{B} = \mu\mu_0\mathbf{H} \quad .. \quad .. \quad .. \quad .. \quad (6)$$

where ϵ_0 and μ_0 are the permittivity and permeability of free space, and ϵ, μ, may be real, imaginary, or complex scalars, or may be tensors whose elements may be real, imaginary, or complex. The only tensors that need be considered are the permeability tensor of a ferrite,

$$\begin{bmatrix} \mu & -j\alpha & 0 \\ j\alpha & \mu & 0 \\ 0 & 0 & 1 \end{bmatrix} \quad .. \quad .. \quad .. \quad .. \quad (7)$$

and the permittivity tensor of a plasma,

$$\begin{bmatrix} \epsilon & -jk & 0 \\ jk & \epsilon & 0 \\ 0 & 0 & 1 \end{bmatrix} \quad .. \quad .. \quad .. \quad .. \quad (8)$$

The conduction current can be regarded as an imaginary part of the displacement $\mathbf{D}$ in a scalar medium, if the time variation is of the form $e^{j\omega t}$, so that $\delta/\delta t = j\omega$. Equ. (3) then becomes

$$\text{Curl } \mathbf{H} = \sigma\mathbf{E} + j\omega\mathbf{D} \quad .. \quad .. \quad .. \quad .. \quad (9)$$

and Equ. (4) becomes

$$\text{Curl } \mathbf{E} = -j\omega\mathbf{B} \quad .. \quad .. \quad .. \quad .. \quad (10)$$

For the right-hand side of Equ. (9) we have $\sigma\mathbf{E} + j\omega\epsilon\epsilon_0\mathbf{E}$; i.e., $j\omega\epsilon\epsilon_0\mathbf{E}$ $(\epsilon - j\sigma/\omega\epsilon_0)$. Replacing ϵ by ϵ', $\sigma/\omega\epsilon_0$ by ϵ'', and writing $\epsilon = \epsilon' - j\epsilon''$, Equ. (9) becomes

$$\text{Curl } \mathbf{H} = j\omega\mathbf{D} = j\omega_0\epsilon\mathbf{E} \quad .. \quad .. \quad .. \quad (11)$$

where $\mathbf{D}$ is now a generalized displacement current, and includes the conduction current. If ϵ is a tensor quantity, the elements of the tensor can also be taken as complex

in order to account for conduction currents. Again, the conductivity σ may itself be a tensor quantity, by virtue of which the dielectric constant becomes a tensor, although the real part of the dielectric constant may be a scalar.

It should be remembered that in any practical medium there will be dielectric losses quite apart from conduction currents. In general, the imaginary part of the dielectric constant includes such losses as well as those due to conduction currents. Magnetic losses also occur, although there is no analogy with conduction currents, and a scalar relative permeability μ can be written $\mu' - j\mu''$ to take account of these. Similarly, in the case of a ferrite, the elements of the permeability tensor can be written $\mu = \mu' - j\mu''$, $\alpha = \alpha' - j\alpha''$.

Any medium can be characterized, for its effects on electromagnetic waves, by appropriate values of permittivity and permeability, and this enables us to generalize the method of deriving and solving a wave equation. It is not necessary to adopt different methods according to the type of medium. The principal types of medium are dielectrics, ferrites, plasmas, and metals or perfect conductors. For a dielectric, we take a scalar relative permeability $\mu = 1$, and a scalar complex relative permittivity $\epsilon = \epsilon' - j\epsilon''$. For a ferrite we take a scalar complex permittivity and a tensor relative permeability whose elements μ and α are complex. For a plasma, the relative permeability is again unity and the relative permittivity is a tensor with, in general, complex elements. A metal is a dielectric with ϵ'' very large, ϵ' small, and μ unity. The perfect conductor, which is usually taken as an approximation to a metal, has ϵ'' infinite, ϵ' zero, and μ zero.

The special values of permeability and permittivity can be conveniently substituted after the wave equation has been solved. In deriving the wave equation, one can start with Maxwell's equations in their most general form, taking either scalar or tensor ϵ and μ but not otherwise taking note of the nature of the medium being treated. When the correct values of ϵ and μ are eventually substituted according to the nature of the medium, the correct results will be obtained. The advantage of this procedure is that it reduces the treatment to a rule of thumb and thereby lessens the likelihood of conceptual errors. Also, the same piece of analysis can frequently be made to serve for more than one problem.

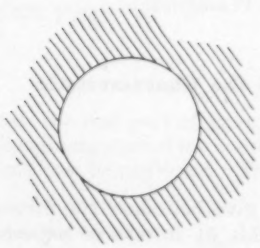

Fig. 4. *Homogenous waveguide. The hatched region, which extends to infinity, represents a perfect conductor, while within the circle any homogenous medium may be present. This medium is a lossless scalar dielectric in the case of the HSP guide*

One simplification of Maxwell's equations that we can make at this point is to put $\rho = 0$; i.e., we shall not be concerned with free charges. Maxwell's equations, in the most general form in which we shall require them, may now be written

$$\left.\begin{aligned} \text{Div } \mathbf{D} &= 0 \\ \text{Div } \mathbf{B} &= 0 \\ \text{Curl } \mathbf{H} &= j\omega\mathbf{D} \\ \text{Curl } \mathbf{E} &= -j\omega\mathbf{B} \end{aligned}\right\} \quad .. \quad .. \quad .. \quad .. \quad (12)$$

We may note in passing that the truth of Equ. (1) is not altered by the modification we have made in **D**, because the divergence of the conduction current must vanish. This follows from the fact that the current must flow in closed loops.[2]

Some simplification is often possible because either the permittivity or permeability, or both, is a scalar quantity in most media. In these cases, we can write Div $\mathbf{D} = \epsilon\epsilon_0$ Div $\mathbf{E}$ and Div $\mathbf{B} = \mu\mu_0$ Div $\mathbf{H}$, so that the first two of Equs. (12) become

$$\left.\begin{aligned} \text{Div } \mathbf{E} &= 0 \\ \text{Div } \mathbf{H} &= 0 \end{aligned}\right\} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (13)$$

But if ϵ or μ is a tensor, the appropriate equation of set (13) does not hold, and Equs. (12) must be used. For this reason, m.k.s. units are to be preferred to c.g.s. because the distinction between inductions and fields is explicit.

The Wave Equation

For each medium in a waveguide, a wave equation can be derived from Maxwell's equations. This equation can be solved for the field equations, and at a boundary surface between two media the field components in the media must satisfy certain conditions. This is a question we shall return to in the next Section (Boundary Conditions and the Characteristic Equation). In the present Section, we are concerned with the wave equations. We shall not give the derivations from Maxwell's equations; these are given elsewhere. Here we shall only give some typical forms of the wave equation that occur.

First we note that the Laplacian operator can be written

$$\nabla^2 = \nabla_t^2 + \frac{\partial^2}{\partial z^2} \quad .. \quad .. \quad .. \quad .. \quad (14)$$

which can be taken as the definition of ∇_t. Thus ∇_t^2 is what remains if, on writing out ∇^2 in full in the appropriate co-ordinate system, $\partial^2/\partial z^2$ is cancelled. Alternatively, ∇_t is the transverse component of ∇, regarded as a formal vector. Wave equations can now be written involving ∇_t^2.

For vacuum, the wave equation is

$$\{\nabla_t^2 + k_0^2\}\left\{\frac{E_z}{H_z}\right\} = 0 \quad .. \quad .. \quad .. \quad (15)$$

and for a homogeneous scalar medium it is

$$\{\nabla_t^2 + k_1^2\}\left\{\frac{E_z}{H_z}\right\} = 0 \quad .. \quad .. \quad .. \quad (16)$$

where $k_0^2 = \omega^2\epsilon_0\mu_0 - \beta^2$ .. .. .. (17)

$$k_1^2 = \omega^2\epsilon_0\mu_0\epsilon\mu - \beta^2 \quad .. \quad .. \quad .. \quad (18)$$

β being the phase constant (i.e., the z variation is taken

to be $e^{-j\beta z}$). ϵ and μ are scalar quantities, but may be real, imaginary, or complex.

For a ferrite, the wave equation is[1]

$$\{(\nabla_t^2 + s_1^2)(\nabla_t^2 + s_2^2)\}\begin{pmatrix} E_z \\ H_z \end{pmatrix} = 0 \quad .. \quad .. \quad (19)$$

where s_1 and s_2 are functions of the properties of the ferrite, of the phase constant β, and of the quantity $\omega^2\epsilon_0\mu_0$. A similar equation to (19) can be written for a plasma, but s_1 and s_2 will then be different functions. This case has been treated by Suhl and Walker[3] and by Hideya Gamo.[4]

Piefke[5] has considered an anisotropic medium with the properties

$$[\epsilon] = \begin{bmatrix} \epsilon_t & 0 & 0 \\ 0 & \epsilon_t & 0 \\ 0 & 0 & \epsilon_z \end{bmatrix} \qquad [\mu] = \begin{bmatrix} \mu_t & 0 & 0 \\ 0 & \mu_t & 0 \\ 0 & 0 & 1 \end{bmatrix} \quad .. \quad .. \quad .. \quad .. \quad (20)$$

No natural medium has such properties; this is an artificial medium made of alternate layers of metal and dielectric, with spacing very much less than a wavelength. The z direction is normal to the planes of the layers. The wave equations for such a medium are

$$\left.\begin{aligned} \{\mu_t\nabla_t^2 + k_t^2\}\, H_z = 0 \\ \{\epsilon_t\nabla_t^2 + \epsilon_z k_t^2\}\, E_z = 0 \end{aligned}\right\} \quad .. \quad .. \quad .. \quad (21)$$

$$\text{where } k_t^2 = \omega^2\epsilon_0\mu_0\epsilon_t\mu_t - \beta^2 \quad .. \quad .. \quad .. \quad (22)$$

Solutions of the Wave Equation

The wave equations in the preceding Section [Equs. (15), (16), (19), and (21)] apply to co-ordinate systems in which there is at least one straight-line axis, which we have taken to be in the z direction. No restriction has so far been imposed on the co-ordinates in the transverse plane, and for the moment we may take them to be ξ and η as in the first part of the second Section (Preliminary Questions). ξ and η should be chosen in such a way that all boundary surfaces can be expressed by simple equations, preferably of the form ξ = constant or η = constant. For example, if all the boundary surfaces are concentric cylinders of circular cross-section, the obvious choice is a circulary-cylindrical co-ordinate system, with $\xi = r$, $\eta = \theta$, and the surfaces are then all of the type r = constant.

If the boundary surfaces are complicated in shape, it may not be possible to choose a convenient co-ordinate system. For a large class of problems, however, this is not the case, and we shall restrict our attention to this class. In all practical problems, it is possible to write the solutions of the wave equations in the forms

$$\left.\begin{aligned} E_z &= f_1(\xi)\cdot g_1(\eta)\cdot e^{-j\beta z}\cdot e^{j\omega t} \\ H_z &= f_2(\xi)\cdot g_2(\eta)\cdot e^{-j\beta z}\cdot e^{j\omega t} \end{aligned}\right\} \quad .. \quad .. \quad (23)$$

Equations for the other field components, E_ξ, E_η, H_ξ, H_η in terms of E_z and H_z can be obtained from Maxwell's equations; in fact, they arise as by-products in deriving the wave equations.

As examples, we take the case of cylindrical co-ordinates, and give the solutions of the various wave equations. The solutions of Equ. (15) are

$$\left.\begin{aligned} E_z &= \{AJ_p(k_0 r) + BY_p(k_0 r)\}\cdot\begin{Bmatrix}\cos \\ \sin\end{Bmatrix} p\theta\cdot e^{-j\beta z}\, e^{j\omega t} \\ H_z &= \{CJ_p(k_0 r) + DY_p(k_0 r)\}\cdot\begin{Bmatrix}\cos \\ \sin\end{Bmatrix} p\theta\cdot e^{-j\beta z}\, e^{j\omega t} \end{aligned}\right\} \quad .. \quad .. \quad (24)$$

The solutions of Equ. (16) are similar, with k_0 replaced by k_1. The solutions of Equ. (19) are[6]

$$\left.\begin{aligned} E_z &= \{AJ_p(s_1 r) + BY_p(s_1 r) + CJ_p(s_2 r) + DY_p(s_2 r)\}\begin{Bmatrix}\cos \\ \sin\end{Bmatrix} p\theta\cdot e^{-j\beta z}\, e^{j\omega t} \\ H_z &= \{\tau_1[AJ_p(s_1 r) + BY_p(s_1 r)] + \tau_2[CJ_p(s_2 r) + DY_p(s_2 r)]\}\begin{Bmatrix}\cos \\ \sin\end{Bmatrix} p\theta\cdot e^{-j\beta z}\, e^{j\omega t} \end{aligned}\right\} \quad .. \quad .. \quad (25)$$

where τ_1 and τ_2 are functions of $\omega^2\epsilon_0\mu_0$, the phase constant β, and the properties of the ferrite or plasma. The solutions of Equ. (21) are

$$\left.\begin{aligned} E_z &= \{AJ_p(k_t r\sqrt{\epsilon_z/\epsilon_t}) + BY_p(k_t r\sqrt{\epsilon_z/\epsilon_t})\}\begin{Bmatrix}\cos \\ \sin\end{Bmatrix} p\theta\cdot e^{-j\beta z}\cdot e^{j\omega t} \\ H_z &= \{CJ_p(k_t r\sqrt{1/\mu_t}) + DY_p(k_t r\sqrt{1/\mu_t})\}\begin{Bmatrix}\cos \\ \sin\end{Bmatrix} p\theta\cdot e^{-j\beta z}\cdot e^{j\omega t} \end{aligned}\right\} \quad .. \quad .. \quad (26)$$

different for E_z and H_z, unlike the media so far considered. They are

In these equations, p is an integer, including zero, and A, B, C, D, are arbitrary constants to be determined by the method described below (Section: Derivation of the Characteristic Equation for a Waveguide).

Boundary Conditions and the Characteristic Equation

Boundary Conditions

It is usual in textbooks to give boundary conditions, for electric or magnetic fields at interfaces between various kinds of media, in different parts of the book, and in different forms, according to the natures of the media. This creates a certain amount of confusion as to what boundary conditions are appropriate to a particular problem by failing to bring out the essential simplicity of electromagnetic phenomena. For a boundary surface between two media of any kind, it is necessary to remember only three facts:

(1) The component of field (electric or magnetic) parallel to the surface is continuous at the surface; i.e., when the surface is approached from either side,

the values of the fields actually at the surface are the same.

(2) The component of induction (electric or magnetic) normal to the surface is continuous at the surface.

(3) No component of field or induction, either electric or magnetic, can be infinite—not even theoretically in an ideal medium.

The application of these rules enables us to cope with any kind of interface. For scalar dielectrics, the treatment is straightforward. Nor is there any difficulty for a ferrite with its tensor permeability, if one remembers to write for the magnetic induction

$$\left.\begin{aligned} B_x &= \mu_0[\mu H_x - j\alpha H_y] \\ B_y &= \mu_0[\mu H_y + j\alpha H_x] \\ B_z &= \mu_0 H_z \end{aligned}\right\} \quad \ldots \quad \ldots \quad \ldots \quad (27)$$

when the polarizing field is in the z direction. A metal can be treated in the same way as a dielectric. For the component of electric field parallel to the surface of the metal we write

$$E_{d\ par.} = E_{m\ par.}$$

where the subscripts d and m refer respectively to the dielectric (or air or vacuum) in contact with the metal, and to the metal. Now, if V is the potential at a point in the surface of the metal, and if $E_{m\ par.}$ is in the x direction, we have

$$E_{m\ par.} = \frac{dV}{dx} \propto \frac{I}{\sigma}$$

I being the current density and σ the conductivity. σ is very high for a metal, so that unless the current density is very high, $E_{m\ par.}$ and therefore $E_{d\ par.}$ must be small. For a perfect conductor, σ is infinite and if I is to remain finite $E_{m\ par}$ and $E_{d\ par.}$ must be zero.

For the normal component of electric induction, we have

$$D_{d\ norm.} = D_{m\ norm.},$$

and if we take the medium d to be free space, this becomes

$$\epsilon_0 E_{d\ norm.} = \epsilon\epsilon_0 E_{m\ norm.}$$

just as for a boundary between free space and dielectric. Now put

$$\epsilon = \epsilon' - j\epsilon'' = \epsilon' - \frac{j\sigma}{\omega\epsilon_0} = -\frac{j}{\omega\epsilon_0}(\sigma + j\sigma'')$$

i.e., the real part of dielectric constant is regarded as an imaginary part of the conductivity. For metals, $\sigma'' << \sigma$, and can be neglected. Thus we have

$$\epsilon_0 E_{d\ norm.} = -\frac{j\sigma}{\epsilon_0\omega} \cdot \epsilon_0 E_{m\ norm.} = -\frac{j\sigma}{\omega} \cdot E_{m\ norm.}$$

Two facts are apparent; first, $E_{d\ norm.}$ is much greater than $E_{m\ norm.}$, so that a large electric field can exist normal to a metallic surface, but does not penetrate the surface, and secondly the electric field in the metal, small though it is, is 90° out of phase with that outside the metal. For a perfect conductor, σ is infinite and $E_{m\ norm.}$ must be zero, while $E_{d\ norm.}$ can still remain finite. $E_{m\ norm.}$ cannot be finite, as this would require $E_{d\ norm.}$ to be infinite; this would mean a wave with infinite energy.

For the magnetic field components, we have

$$H_{d\ par.} = H_{m\ par.}$$

and

$$B_{d\ norm.} = B_{m\ norm.}$$

For most metals, in particular copper, the relative permeability is 1; we shall not consider ferromagnetic metals and alloys, which are not used as materials for waveguides. Thus

$$B_{par.} = \mu_0 H_{par.}$$

and

$$B_{norm.} = \mu_0 H_{norm.}$$

both inside and outside the metal. However, if a perfect conductor is taken as an approximation to a metal, it is necessary to take into account the relationship between **E** and **H** imposed by Maxwell's equations. For a perfect conductor, $\mu=0$, so that $B_{par.}$ and $B_{norm.}$ must be zero. This does not affect the relationship $H_{d\ par.} = H_{m\ par.}$, so that it is still possible to have a finite parallel component of magnetic field. For the normal component, we write $B_{d\ norm.} = \mu_0 H_{d\ norm.}$ and $B_{m\ norm.} = \mu\mu_0 H_{m\ norm.}$; i.e., $H_{d\ norm.} = \mu H_{m\ norm.}$ Since $\mu=0$, $H_{d\ norm.} = 0$; i.e., the component of magnetic field externally normal to the surface of a perfect conductor vanishes on the surface. $H_{m\ norm.}$, however, is not necessarily zero.

Derivation of the Characteristic Equation for a Waveguide

We have now obtained solutions for the field components as functions of the properties of the medium, position, the phase constant β, and certain arbitrary constants. We also know the physical conditions that must be satisfied at a boundary surface between two media. The next steps are to evaluate β and to use this value to determine the arbitrary constants. It will be noticed that these occur linearly, i.e., to the first power only, in the field equations. Also, the boundary conditions involve the field components or their derivatives only linearly. Thus, when the field functions for the various field components are substituted for the latter in one of the boundary conditions, an equation results that is linear in the arbitrary constants. An example of this will be given in the second article of this series.

Further, in this equation no term occurs that does not contain an arbitrary constant; such an equation is said to be homogeneous in the arbitrary constants – all the terms are of the same degree (the first). Suppose that in the field equations for a given waveguide there are n arbitrary constants. Then it is possible to divide through by any one of them to obtain a non-homogeneous equation in n-1 ratios of arbitrary constants. By applying n-1 boundary conditions, a set of equations is obtained which could be solved for these n-1 ratios, if the coefficients were known. The coefficients are functions of the positions of the boundary surfaces, and of the properties of the media; these are known or can be stipulated. They are also functions of β, which has still to be determined.

Another boundary condition is now taken. Thus we have n equations for n-1 unknown ratios or, if the division is not performed, a set of n linear homogeneous equations in n unknowns – the arbitrary constants. The condition that these equations are mutually consistent is that the determinant of the coefficients vanishes. This can be

seen by noting that if the set is consistent, any one equation can be expressed as a linear sum of all the others, while if in a determinant one row or column can be expressed as a linear sum of the others the determinant vanishes. The determinant equated to zero gives an equation in which the only unknown is β. This equation is the characteristic equation. Once β has been obtained from this, the arbitrary constants are obtained quite simply from the set of linear equations, and so the field components are known.

So far, the technique is straightforward, and if a suitable model is chosen there is no great difficulty in obtaining a characteristic equation. Sometimes the model may not be a very good approximation to the physical situation, which immediately limits the value of the treatment, although in most cases this need not happen. An example of a case where the model falls short of what is ideally desirable is the disc line treated by Piefke[5], in which a medium consisting of alternate thin layers of metal and dielectric is regarded as an anisotropic homogeneous medium. A reasonable approximation is obtained in this way, but features of the guide which are dependent on the periodicity of the structure will not be detected.

For a HSP guide the characteristic equation is usually fairly simple; for example, if the guide is of circular cross-section, it is of the form

$$J_p(ka) \, . \, J_p'(ka) = 0 \quad .. \quad .. \quad .. \quad .. \quad (28)$$

where p is an integer (including zero), a is the radius of the guide, and $k^2 = \omega^2\epsilon_0\mu_0\epsilon\mu - \beta^2$, ϵ and μ being the relative permittivity and permeability of the material of the guide. Either factor on the left-hand side can be equated to zero, giving

$$J_p(ka) = 0 \quad .. \quad .. \quad .. \quad .. \quad (28a)$$

for E modes, and

$$J_p'(ka) = 0 \quad .. \quad .. \quad .. \quad .. \quad (28b)$$

for H modes. At given frequency, these equations can be solved quite easily for β. The cut-off equations are obtained by putting $\beta = 0$, giving

$$\left.\begin{aligned} J_p\,(\omega a\sqrt{\epsilon_0\mu_0\epsilon\mu}) &= 0 \\ \text{and } J_p'(\omega a\sqrt{\epsilon_0\mu_0\epsilon\mu}) &= 0 \end{aligned}\right\} \quad .. \quad .. \quad .. \quad (29)$$

For a guide of given radius, Equs. (29) give the cut-off frequencies for the various modes. A mode is defined, in the present instance, by assigning a value to p, and taking the qth solution of Equ. (28a) or (28b); this gives the phase constant of the E_{pq} or H_{pq} mode respectively.

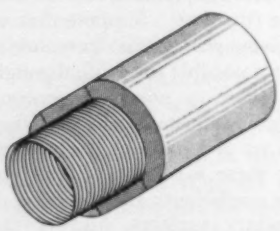

Fig. 5. *Helix guide*

A characteristic equation is usually difficult to solve, and the methods used are outside the scope of this general discussion. It is the solutions that are of interest to the microwave scientist, rather than the means of getting them. The difficulty increases as the guide departs further from the HSP guide. Thus if lossy media are treated; if more boundary surfaces are included; if a scalar medium is replaced by a tensor medium; or if a medium is included whose properties are functions of position; any of these steps, or any combination of them, renders the characteristic equation more complicated. There is great interest nowadays in complicated waveguides, and many types await detailed study. For two particularly important cases, however, detailed studies of the mode spectrum have been made. These are the guide whose wall is formed of a helically-wound wire, regarded as an infinitesimally thin anisotropic conducting surface, the whole being surrounded by a lossy dielectric[7] (Fig. 5); and the guide of circular cross-section, bounded by a perfect conductor, and containing a longitudinally-magnetized concentric rod of ferrite of arbitrary radius[1] (Fig. 6).

Some simplification of the characteristic equation is sometimes obtained by putting $\beta = 0$ to get a cut-off condition. It is usually not possible to split the characteristic equation proper into two factors (the characteristic equation for the HSP guide is an exceptionally simple case), but the cut-off equations often do split into two factors, which greatly simplifies the task of deciding the mode nomenclature. For it is found that the modes given by one of the pair of cut-off equations are related to the H modes of the HSP guide, while those given by the other cut-off equation are similarly related to the E modes of the HSP guide. Further, solution of the cut-off equations tells us, for a given set of parameters (geometry of the system, properties of the media), which modes are able to propagate, and this information serves as a starting point for an attempt to solve the characteristic equation proper. This method was applied to the problem of the guide of circular cross-section with ferrite[1]. It fails, however, with the problem of a guide of rectangular cross-section containing a transversely-magnetized slab of ferrite, because the cut-off equation does not split into two factors and remains almost as difficult to solve as the characteristic equation proper. It is partly for this reason that a detailed study of this guide has not so far been carried out, although the H_{10} mode has been treated[8,9]. A more general, but less detailed, discussion has been given by Barzilai and Gerosa[10].

The Characteristic Equation for a Cavity

The characteristic equation for a cavity which consists of a finite length of waveguide terminated by perfectly-conducting surfaces normal to the z axis is identical in form with that of the waveguide. However, the wavelength in the z direction is now determined by the length l of the cavity, so that the characteristic equation is now not to be solved for β. The wavelength is $2\pi/\beta$, and must be such that the length l is an integral number of half-wavelengths. Thus

$$\frac{2\pi}{\beta} = \frac{2l}{n} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (30)$$

where n is an integer. The modes are now characterized by three integers, p, q, and n. In the characteristic equation, β is replaced by $n\pi/l$, and the properties of

the relevant media and the geometry are given. Any frequency may be fed to the cavity, but the cavity will only respond if the frequency satisfies the characteristic equation. Values of p and n must be substituted, and there will then be an infinite number of permissible values of frequency, in two series. The qth value in one series corresponds to the H_{pqn} mode, the qth value in the other series to the E_{pqn} mode.

If $n = 0$, $\beta = 0$, and the characteristic equation becomes identical with the cut-off equation of the corresponding waveguide. In both cases, all the other parameters being given, it is the frequency that is to be solved for, so that to study the cut-off conditions in a waveguide is mathematically – and, virtually, physically – the same problem as to study the E_{pqo} and H_{pqo} modes of the corresponding cavity.

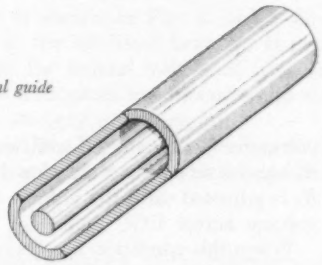

Fig. 6. *Ferrite-loaded cylindrical guide*

As an example of this, Heller[11] and Bussey and Steinert[12, 13] have given exact treatments of the H_{pqo} modes of a cavity of circular cross-section containing a concentric ferrite rod, and give a characteristic equation which is equivalent to the H-mode cut-off equation for the corresponding waveguide, given in reference 1.

Conclusion

A scheme has been described for obtaining the characteristic equation for any waveguide, or any cavity that can be regarded as a truncated waveguide. The solution of this characteristic equation is usually very difficult and is not discussed. Some simplification of the study can sometimes, but not always, be made in the waveguide case by making a preliminary study of the cut-off equation.

As well as giving some understanding of the mathematician's task, some of the difficulties with which he has to contend are pointed out.

The application of these principles to certain problems will be treated in the second article in this series, and it will be shown how a more general approach can often shed light on the behaviour of a system which is usually treated on the basis of a simple model. Such questions will be discussed as the transmission-line mode from the waveguide point of view, and the apparent non-existence of certain modes in waveguides. In fact, the non-existent modes are not so non-existent as is generally thought, and they are related to the transmission-line modes of two-conductor guides.

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REFERENCES

1. R. A. Waldron, "Electromagnetic Wave Propagation in Cylindrical Waveguides Containing Gyromagnetic Media", *J. Brit. Instn Radio Engrs*, 1958, Vol. 18, pp. 597–612, 677–690, 733–746.
2. See, e.g., B. Hague, "An Introduction to Vector Analysis", Methuen's Monographs, 1946, p. 53.
3. H. Suhl, L. R. Walker, "Topics in Guided Wave Propagation through Gyromagnetic Media", *Bell Syst. tech. J.*, 1954, Vol. 33, pp. 579–659, 939–986, 1133–1194.
4. Hideya Gamo, "The Faraday Rotation of Waves in a Circular Waveguide", *J. phys. Soc. Japan*, 1953, Vol. 8, pp. 176–182.
5. G. Piefke, "Wellenausbreitung in der Schieben-Leitung" (Wave Propagation in the Disc Line), *Arch. elekt. Übertagung*, 1957, Vol. 11, pp. 49–59.
6. R. A. Waldron, "Theory of the Mode Spectra of Cylindrical Waveguides Containing Gyromagnetic Media", *J. Brit. Instn Radio Engrs*, 1959, Vol. 19, pp. 347–356.
7. S. P. Morgan and J. A. Young, "Helix Waveguide", *Bell Syst. tech. J.*, 1956, Vol. 35, pp. 1347–1384.
8. B. Lax, K. J. Button and L. M. Roth, "Ferrite Phase Shifters in Rectangular Waveguide", *J. appl. Phys.*, 1954, Vol. 25, pp. 1413–1421.
9. R. F. Soohoo, "Cut-off Phenomena in Transversely Magnetised Ferrites", *Proc. Inst. Radio Engrs*, 1958, Vol. 46, pp. 788–9.
10. G. Barzilai, G. Gerosa, "Modes in Rectangular Guides Filled with Magnetised Ferrite", *Nuovo Cim.*, 1958, Vol. 7, pp. 685–697.
11. G. S. Heller, "Ferrite-Loaded Cavity Resonators", paper presented at the International Congress on Microwave Circuits and Aerials, Paris, October 1957.
12. H. E. Bussey, L. A. Steinert, "An Exact Solution for a Cylindrical Cavity Containing a Gyromagnetic Material", *Proc. Inst. Radio Engrs*, 1957, Vol. 45, pp. 693–4.
13. H. E. Bussey, L. A. Steinert, "Exact Solution for a Gyromagnetic Sample and Measurements on a Ferrite", *Trans. Inst. Radio Engrs*, MTT6, 1958, pp. 72–76.

MANUFACTURERS' LITERATURE

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Measurement of Impedance at Audio-Frequency

MODULUS AND ANGLE WITH SIMPLE APPARATUS

By N. P. Scholes*, A.M.I.E.E. and J. E. Macfarlane*, B.Sc.(Eng.), M.I.E.E., A.M.I.Mech.E.

The measurement of electrical impedance and phase angle by conventional a.c. bridge methods has certain disadvantages when the measurement is required over a range of frequency and particularly if this involves a change in the sign of the phase angle. The bridge method will necessitate the use of a suitable standard of inductance or capacitance which may not always be available, the bridge may be tedious to balance and lengthy calculations are often needed. Difficulties may also arise

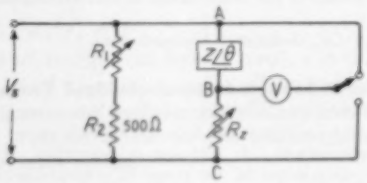


Fig. 1. *Determination of modulus*

in the choice of a suitable detector as the frequency changes. For these reasons other methods may be preferred where a very high degree of accuracy is not required.

The measurement of the modulus of an impedance by means of standard resistors and a voltmeter is well known and can readily be made independent of the accuracy of the instrument, but the determination of the phase angle generally depends on the instrument being correct and, unless its calibration is known, the accuracy will not usually approach that obtainable on an a.c. bridge. A method of determining both modulus and angle of an impedance independent of instrument accuracy is described below and employs only standard variable resistors. The measurements can be made more rapidly than on an a.c. bridge and the accuracy compares favourably with many commercial bridges. Direct reading for both modulus and power factor are obtained, hence avoiding the usual calculations for these parameters.

Determination of Modulus |Z|

The circuit diagram is shown in Fig. 1 in which R_2 is a fixed resistor of 500 ohms and R_1 and R_x are variable standard resistors; V_s is a stabilized a.c. source. The voltmeter V must be of sufficiently high resistance, and R_1 can be set at any value for this first test. The resistor R_x is adjusted until the voltage across AB is equal to the voltage across BC; i.e., $|V_{ab}| = |V_{bc}|$ then $R_x = |Z|$.

When this condition has been satisfied it is convenient to adjust V_s so that V_{ab} is about three-quarters of the full-scale reading of the voltmeter and the source voltage is maintained constant.

Determination of the Angle θ

The circuit diagram is shown in Fig. 2 in which the voltmeter connection is altered but R_x remains at the value obtained in the first test.

The resistor R_1 is adjusted until the p.d. between E and B is equal to the p.d. across R_2; i.e., $|V_{eb}| = |V_{ec}|$

For this condition

$$\cos\theta = \frac{R_1 - R_2}{2R_2}$$

which is the power factor, hence θ can be evaluated.

The derivation of this expression is straightforward. So that cos θ may be read directly it is convenient to make $R_1 = R_A + R_2$ whence R_A is variable.

$$\text{Then } \cos\theta = \frac{R_A + R_2 - R_2}{2R_2}$$

$$\text{i.e., } \cos\theta = \frac{R_A}{1000}$$

The significant figures for cos θ are now obtained directly on R_A.

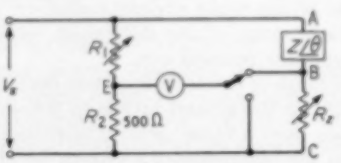


Fig. 2. *Determination of angle*

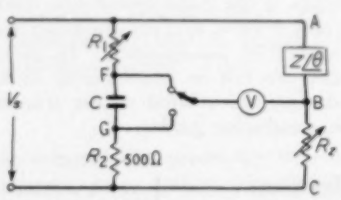


Fig. 3. *Determination of the sign of θ*

* College of Technology, Liverpool.

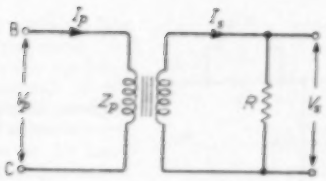


Fig. 4. *Transformer test*

Determination of the Sign of the Angle θ

The circuit diagram is shown in Fig. 3 which now includes a capacitor C in the left-hand branch. It may be desirable to arrange for several values of C to be switched into circuit, as difficulties may occur when the two measured voltages are very nearly equal.

In this test if V_{bf} is less than V_{bg} then θ is positive and, conversely, when V_{bf} is greater than V_{bg} θ is negative.

Example

As an example the determination of the efficiency of a small audio-frequency transformer may be considered.

Direct measurements of input and output power are inaccurate due to the small values of power, as suitable wattmeters are not generally available.

Consider the simple circuit shown in Fig. 4 in which the load on the transformer is an accurately known resistor R.

Input power $= V_p I_p \cos\theta$, where $\cos\theta$ is the power factor.

Output power $= \dfrac{V_s^2}{R}$ hence

the efficiency $\eta = \dfrac{V_s^2}{R\,V_p I_p \cos\theta}$

$= \dfrac{V_s^2\,Z_p}{R V_p^2 \cos\theta}$

Z_p and $\cos\theta$ are determined independently of the voltmeter accuracy and R is known, hence η is dependent on the ratio V_s^2/V_p^2.

Provided the same voltmeter is used to measure these two voltages and assuming a uniform percentage error in the instrument, then no appreciable inaccuracy occurs.

Correspondence

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Angle-Modulated Waves

Sir,—In his two articles on "Power in Angle-Modulated Waves" (*Electronic & Radio Engineer*, August 1959, and *Electronic Technology*, October 1960), W. C. Vaughan treats us to many pages of mathematics but, I fear, misses the essential point concerning the mean (or r.m.s.) value of a varying function.

If a function is periodic in time T, its mean value is calculated over an interval T. If a function is not periodic, it has no mean value, though in practice $\int_{t_1}^{t_2} f(t)dt/(t_2-t_1)$ usually tends to a limit as (t_2-t_1) tends to infinity, and this limit is then called the 'mean value' of $f(t)$.

Dr. Vaughan considers the mean value of $\cos^2(\omega t + \theta\cos pt)$ which, by expanding into $(\frac{1}{2} + \frac{1}{2}\cos 2\ldots)$, amounts to considering the mean of $f(t) = \cos(2\pi t/T_1 + \phi\cos 2\pi t/T_2)$, where $2\pi/T_1 = 2\omega$, $\phi = 2\theta$ and $2\pi/T_2 = p$.

The function $f(t)$ is periodic if T_1/T_2 is rational, and its periodicity T is then the L.C.M. of T_1 and T_2.

If T/T_2 is even (implying T/T_1 odd), we see at once that the mean of $f(t)$ is zero since $f(t + T/2) = -f(t)$.

For T/T_2 equal to an odd number greater than 1, we only need to recall that f can be analysed by well-known methods into a centre-frequency component and an infinite sequence of side-frequency components. Since all these are sinusoids, their mean (over interval T) $= 0$, and so the mean of f is likewise zero.

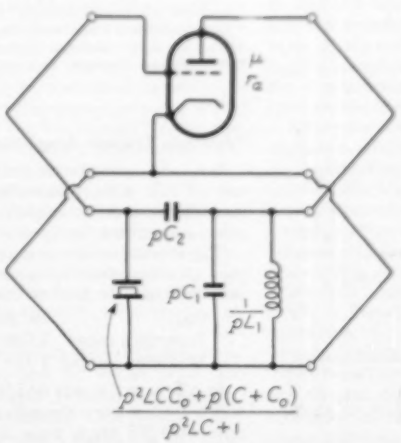
If $T = T_2$, then one of the side-frequency components has exactly zero frequency and therefore represents a d.c. level equal to the amplitude of this particular component. This then is the mean of $f(t)$.

A. L. Robson.

London, N.19.
30th January 1961.

Determinants in Theory of Harmonic Oscillators

Sir,—In his letter in the December issue, Mr. J. W. Head recommends the "full use" of 2-port matrices in the analysis of Mr. Kitai's oscillator. Why, therefore, does he not proceed as indicated below?



$$Y_1 = \begin{bmatrix} 0 & 0 \\ \frac{\mu}{r_a} & \frac{1}{r_a} \end{bmatrix}$$

$$Y_2 = \begin{bmatrix} \frac{p^2LCC_0+p(C+C_0)}{p^2LC+1}+pC_2 & -pC_2 \\ -pC_2 & p(C_1+C_2)+\frac{1}{pL_1} \end{bmatrix}$$

$p = j\omega$

For oscillation the determinant of the matrix sum $\mathbf{Y}_1 + \mathbf{Y}_2$ must be zero. This gives the result required by Mr. Kitai.

S. R. Deards.

Department of Aircraft Electrical Engineering,
The College of Aeronautics,
Cranfield.
10th January 1961.

The Junction Transistor

Sir,—Although the junction transistor is most commonly used in the earthed-emitter connection (corresponding to the commonly-used earthed-cathode configuration for valves), both the explanation of its operation as well as the basic equivalent circuit are usually still given in terms of the earthed-base structure. This is undoubtedly a 'hangover' from the point-contact transistor and it is high time that this practice was given up. Recently, a number of workers[1,2,3,4] have pointed out the necessity of using the earthed-emitter structure as the basic one. As this is the most commonly used configuration in transistor circuits and usually gives maximum voltage and power gain, it can be considered as the basic configuration in its own right. Also, an equivalent circuit and its explanation in terms of earthed-emitter structure will make the similarities (rather dualities) between transistor and valve circuits more evident. This will lead to better understanding and easier design of transistor circuits.

The basic difference between a transistor and a triode valve is that the former is a current-operated device whereas the latter is a voltage-operated device. The electrons flowing from cathode to anode in a vacuum triode are controlled by the grid voltage. But the minority carriers (electrons or holes, as the case may be) flowing from emitter to collector in a transistor are controlled by the base current. Now the grid structure in a vacuum triode is a mesh of wire with large openings, and the accelerating voltage at the anode is able to cause a forward field (for electrons) at the cathode even when the grid is reverse (negatively) biased. But the situation is a bit different in a transistor where the base region is a solid layer of semiconductor having very little openings in the crystal lattice. Thus, it is not possible for the accelerating voltage at the collector (with respect to the minority carriers injected in the emitter) to exert a forward field in the emitter when the base is reverse biased. Hence having the base forward biased (corresponding to positive grid operation in a valve) becomes a 'necessary evil' for the operation of the transistor; i.e., for the minority carriers to flow from emitter to collector.

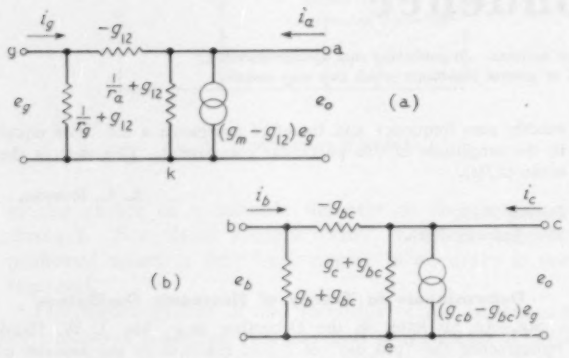


Fig. 1. *Basic equivalent circuits of (a) positive-grid triode in earthed-cathode configuration and (b) transistor in earthed-emitter configuration*

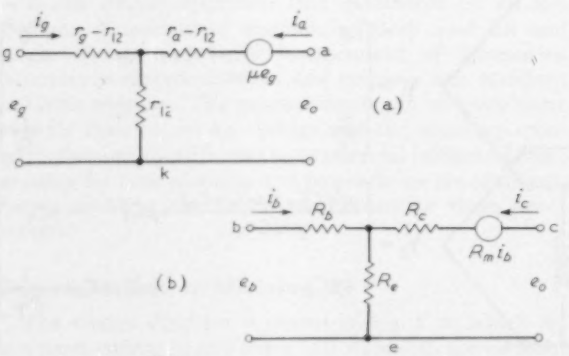


Fig. 2. *Equivalent T-structure of the circuits shown in Fig. 1: (a) positive-grid triode, (b) transistor*

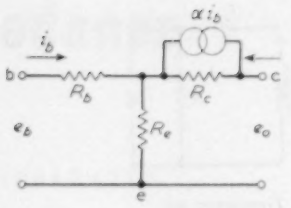


Fig. 3. *Constant-current equivalent form of the transistor equivalent circuit in Fig. 2(b)*

As in a vacuum triode, this forward biasing in a transistor is associated with the consequent disadvantages of low input impedance and increased feedback. But, unlike the triode valve, we have no better choice (such as, using a reverse-biased base). Otherwise a vacuum triode and a transistor have similar operating characteristics and the equivalent circuits of a positive-grid triode and a transistor are almost identical as discussed in the next section.

Fig. 1 (a) shows the low-frequency a.c. equivalent circuit of a triode (earthed-cathode) with positive-grid operation; r_g and r_a are respectively the grid and anode resistances and g_{12} gives the change in grid current for a unit change in anode voltage and is negative. Fig. 1(b) shows the corresponding equivalent circuit for a transistor in the earthed-emitter configuration. The similarities are evident. Figs. 2(a) and (b) show respectively the equivalent T-structure of triode and transistor circuits as shown in Fig. 1. As a transistor is a current-operated device, the equivalent circuit would be better represented by a constant-current generator as shown in Fig. 3. Here the current amplification factor $\alpha \gg 1$ (corresponding to voltage amplification factor $\mu \gg 1$ in a triode). As it is a current-operated device, the collector characteristics are more like the anode characteristics of a vacuum pentode rather than a triode. The other transistor configurations like earthed-base and earthed-emitter can easily be derived from this basic structure.

Thus for the earthed-base, $\alpha_b = \dfrac{\alpha}{1+\alpha}$ and $r_c = R_c\,(1+\alpha)$.

B. Chatterjee.

Communication Engineering Dept.,
Indian Institute of Technology,
Kharagpur, India.
14th January 1961.

REFERENCES

[1] Editorial, *Electronic Technology*, June 1960, p. 213.
[2] M. K. Achuthan, *Electronic Technology*, June 1960, Vol. 37, p. 238.
[3] "Cathode Ray", *Electronic Technology*, August 1960, Vol. 37, p. 323.
[4] B. Chatterjee, *Electronic Technology*, December 1960, Vol. 37, p. 475.

Multiple Transit Amplification by Travelling-Wave Tube

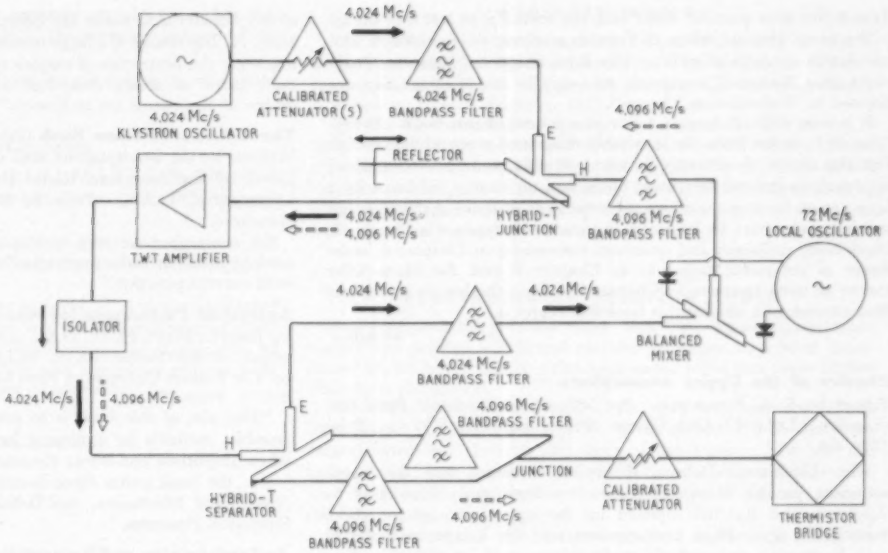
Sir,—An investigation has been made of the possibility of making use of the wide bandwidth of the microwave travelling-wave amplifier in order to amplify a fairly narrow-band signal more than once, each transit being at a slightly different frequency.

The investigation was conducted on a CV2188 transmitting tube, part of a Standard Telephones and Cables repeater station. The nominal characteristics of this tube are:

Gain	16–20 dB (gain decreases as signal increases)
Frequency range	3,600–4,200 Mc/s
Maximum output	2 W.

As a first step it was decided to attempt to pass a signal of bandwidth 1 Mc/s twice through the amplifier, once at 4,024 Mc/s and once at 4,096 Mc/s, these frequencies being chosen to suit com-

ponents already available. The circuit which was finally developed is shown in the figure (right).



The input signal at 4,024 Mc/s from a reflex klystron oscillator passes to the t.w.t. through the hybrid-T junction, the spare side arm of this junction being terminated in a short-circuit so positioned that all the energy, both at 4,024 and 4,096 Mc/s is routed to the t.w.t. input. After the first transit the amplified 4,024-Mc/s signal is routed to the balanced mixer via the hybrid-T separator. In this component separation of the two frequencies is secured by a suitable choice of distances of the two 4,096-Mc/s bandpass filters from the junction. The balanced mixer uses two CV364 crystals switched in antiphase at 72 Mc/s by the output of the local oscillator. Of the two sidebands generated, only the upper at 4,096 Mc/s is selected and this is passed via the hybrid-T junction to the t.w.t. for the second transit, thence via the separator to the load.

Successful double-pass amplification has been achieved but with an overall gain of only 10 dB. The poor gain is attributed to the t.w.t., an old one, which has been giving a gain of only 12·5 dB per transit, and to the fact that time has not permitted proper tuning of the hybrid-T junctions. The optimum mixer conversion loss is 6 dB and it is thought that with attention to the two matters just mentioned a gain of 23 dB should be possible. Once this has been established it is intended to re-engineer the circuit for four transits.

This investigation was carried out by the following students of the Technical Staff Course as their final design study: Captain L. J. O'Brien, R.C.E., Major J. G. Cochran, R.A., Captain R. A. Bell, R.A., and Captain J. A. S. Walker, R.A.S.C. Thanks are due to Messrs. Standard Telephones and Cables for supplying the transmitting bay of one of their repeater stations, to Messrs. Pye for supplying the microwave gear and to A./Prof. Potok and the staff of the Radar and Telecommunications Branch of the Royal Military College of Science for their valuable assistance.

D. A. James

Royal Military College of Science,
Shrivenham, Wilts.,
3rd January 1961.

Linear Amplifier for Decimicrosecond Pulses

Sir,—The approach proposed to the problem of pulse amplifiers in your September 1959 issue seems to us open to some fundamental objections.

The authors of this paper propose to discard well-known negative-feedback techniques in favour of a method in which a separate amplifier (V_2, V_3 in Fig. 2 of the paper) provides additional gain in order to compensate for the loss of high frequencies in the output stage. Actually, V_2 and V_3 are a positive-feedback pair operating on the brink of oscillation ($k = 1$).

The main defect of this method is the very critical initial-gain adjustment by R_1 which is required to make k very nearly equal to one. Furthermore, every change in external loading or even variation of the h.t. supply, heater voltage, valve characteristics, as well as of the components, requires a re-adjustment of R_1. It should further be pointed out that the linearity of this amplifier cannot be any greater than that of the g_m of V_1 as a function of signal parameters.

We feel that the method is inferior to negative-feedback ones, such as the shunt-regulated amplifier, and we would point out that these can also be arranged to handle positive and negative transients by driving a valve into conduction. Consequently, the quiescent current required need still be only a fraction of the peak current which the output valves are capable of delivering into a capacitive load.

A. F. Fischmann-Arbel.
I. Bar-David.

Ministry of Defence,
Tel Aviv, Israel.
26th February 1960.

[Owing to a misunderstanding, publication of this letter has been delayed. Ed.]

New Books

Electrical Noise

By W. R. Bennett. Pp. 280 + viii. McGraw Hill Publishing Co. Ltd., 95 Farringdon Street, London, E.C.4. Price 77s. 6d.

The author has long been associated with communication problems at the Bell Telephone Laboratories, and this is reflected in the inclusion in the book of noise weighting curves both for audible interference and for visual interference on television. He is best known for work on noise in the popular sense (i.e., in or near the audio-frequency range), but he points out that the use of electrical techniques has wiped out any sharp boundary between acoustic noise and the more general sense of the term. The author sets out to lead the reader from familiar examples of random phenomena and the properties of circuits to the more general characteristics of noise, in the earlier stages carefully avoiding abstract ideas like the equivalent spectrum of shot noise. Thermal noise is introduced via the fundamental principle of equipartition, though Nyquist's transmission-line derivation is not described.

After reviewing the thermal-noise properties of circuits, the book explains carefully the nature of distribution functions, with examples from quantizing noise and ignition noise as well as the more usual gaussian, Rayleigh and Poisson distributions. Shot noise in thermionic valves is introduced via the Llewellyn-Peterson coefficients, and the noise contributions from current partition in multi-electrode valves, induced grid noise, etc., are described; and the following chapter deals with noise in semiconductor devices.

We then return to the properties of black-body radiation, and the origin of Planck's quantum theory, as an introduction to the noise properties of masers which depend on the exchange of energy between radiation and resonators. Chapters on "Noise-generating Equipment" and "Noise Measurements and Techniques" precede the chapter on low-noise equipment, which includes a sketchy treatment of electron-beam valves but a good treatment of non-linear-reactance (parametric) amplifiers, and culminates in the

Haus-Adler *noise measure*. After this, the reader is at last introduced to the more abstract ideas of Fourier analysis, power spectra and correlation analysis of noise. The final chapter, "Noise in Communication Systems", comprises 46 pages in all, of which five are devoted to "Information Theory".

It is very difficult to give a fair assessment of this book. Everything in it (apart from the inevitable misprints) is sound and useful. But the depth of coverage is very variable and the 'concentric' approach (a general treatment of various topics first, followed by a later return for a more advanced treatment of some of them) is not to everyone's taste: for example, thermal noise appears in Chapter 2, black-body radiation and quantum corrections in Chapter 6, noise figure of composite networks in Chapter 8 and the Haus-Adler theory of noise measure in Chapter 9. None the less, it is unlikely that anyone who obtains this book will regret it.

D.A.B.

Physics of the Upper Atmosphere

Edited by J. A. RATCLIFFE. Pp. 586 + xi. Academic Press Inc. (London) Ltd., 17 Old Queen Street, London, S.W.1. Price 103s. 6d.

The Lindemann-Dobson theory of meteors and the echo-sounding in the Heaviside Layer by Breit and Tuve and by Appleton and Ratcliffe opened up the upper atmosphere about forty years ago. High temperatures and the E-layer and the F_1 and F_2 layers; troposphere, stratosphere, and ionosphere; these must have seemed until fairly recently to represent the ultimate in attainable knowledge. This volume, with its account of work that has been chiefly done during the past ten years, shows that we are in fact only beginning to explore the upper atmosphere, and gives a detailed survey of many of the main fields of work. It is written primarily for the research worker in one field, to keep him posted about what has been going on in the others, and is a collation of monographs by different specialists.

Sydney Chapman deals with the earth's outermost atmosphere–the thermosphere–and its structure above the 80-km level as it tails off first to neutral hydrogen and then to ionized hydrogen, the protosphere or geocorona. M. Nicolet follows with an account of the properties and constitution of the atmosphere at these levels, mentioning the conclusions that can be drawn from the acceleration of satellites about fluctuations in density. H. E. Newell discusses satellite and rocket observations and the use of mass spectrometers to determine the compositions. H. Friedman describes observations on the sun's ionizing radiations, in particular, studies of X-rays in the region of about 7 A.U. which present formidable instrumental difficulties. D. R. Bates has three chapters on the airglow and the aurora. J. A. Ratcliffe and K. Weekes describe recent work on the ionosphere. The relation between geomagnetism and other physical phenomena, with particular reference to the upper atmosphere, is discussed by E. H. Vestine. H. G. Booker gives an account of radar studies of the aurora. And J. S. Greenhow and A. C. B. Lovell describe the radar techniques for studying the trails of meteors in the upper atmosphere, which almost takes us back to the beginning of things with its combination of meteor studies and echo-sounding.

The book was written during the period of the International Geophysical Year, 1957–58, and some of the main advances during this period are dealt with in a final chapter, which brings the story up to the end of 1959. It is excellently produced, and it seems churlish to cavil at the price. But the various authors have acquitted themselves so well and so clearly that the book could well be recommended to librarians as a reference book for the ordinary scientific reader with no specialist acquaintance with the subject who wants to know what has been going on up there during the past decade–were it not so expensive. G.R.N.

Sound Absorbing Materials

By E. J. EVANS, B.Sc. and E. N. BAZLEY, B.Sc. Pp. 49. Published for D.S.I.R. by Her Majesty's Stationery Office, York House, Kingsway, London, W.C.2. Price 3s. (by post 3s. 4d.), U.S.A. 54 cents.

"The National Physical Laboratory has been engaged for many years in the determination of the sound absorption coefficients of materials . . . A large number of proprietary materials have been tested on behalf of industry, and the acoustic effects of special treatments and mountings have been investigated. The measurements cover most of the absorbents used in this country, and the purpose of this booklet is to make the information obtained generally available . . . The results of a large number of measurements are collected together; the properties of various types of absorbents are reviewed, their mode of action described and recent developments in the design of absorbents are indicated."

The A.R.R.L. Antenna Book (9th Edition)

Written by the headquarters staff of the A.R.R.L. Pp. 320. Published by the American Radio Relay League, West Hartford 7, Connecticut, U.S.A. Price $2·00 in the United States, $2·25 elsewhere.

"A comprehensive text on the theory and practice of antennas used by amateur radio operators, fully revised and expanded in line with current practice."

Analytical Techniques for Non-Linear Control Systems

By JOHN C. WEST, Ph.D., D.Sc., A.M.I.E.E., A.M.A.I.E.E. General editor, Arthur Porter, Ph.D., M.I.E.E. Pp. 223 + xii. Published by The English Universities Press Ltd., 102 Newgate Street, London, E.C.1. Price 25s.

"The aim of this book is to present, in as simple a manner as possible, methods for analysing feedback control system behaviour when amplitude non-linear elements dominate the performance.

. . . the book unites three broad techniques: the Phase Plane of Non-Linear Mechanics, the Describing Function and the Use of Stochastic Processes."

An Introduction to Electrotechnology

By S. J. KOWALSKI, Dipl.Ing., A.M.I.PolishE. Pp. 301 + xiv. Published by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 35s.

"This work covers approximately the requirements of Part One: Applied Electricity, for the London B.Sc.(Eng.) degree. It is based on lecture courses given by the author. . . .

The discussion of electric and magnetic fields is more detailed than is usual in introductory tests, and the development of electromagnetic theory has been based to some extent on logical arguments to provide a useful physical background to the analysis of electromagnetic phenomena."

The Power Gain of Multi-Tiered V.H.F. Transmitting Aerials

By P. KNIGHT, B.A., A.M.I.E.E. and G. D. MONTEATH, B.Sc., A.Inst.P., A.M.I.E.E. Pp. 38. B.B.C. Publications, 35 Marylebone High Street, London, W.1. Price 5s.

A method of calculating the power gain of multi-tiered v.h.f. transmitting aerials is described in this monograph. Results computed for a comprehensive range of variables are presented in the form of tables.

Semiconductor Technology

Edited by G. K. T. CONN, M.A., PH.D. Pp. 74 + v. Butterworths Scientific Publications, 4 & 5 Bell Yard, London, W.C.2. Price 10s. 6d. (postage 7d. extra).

"A series of papers revised and reprinted from *Research*, Vol. 11, 1958 and Vol. 12, 1959."

Basic Mathematics for Electronics (2nd Edition)

By NELSON M. COOKE. Pp. 679. Published by McGraw-Hill Publishing Co., 95 Farringdon Street, London, E.C.4. Price 58s.

"This is the second edition of the textbook originally entitled 'Mathematics for Electricians and Radiomen'."

Selected Semiconductor Circuits Handbook

Edited by SEYMOUR SCHWARTZ. Eleven parts, pp. 483 + xiv. Published by John Wiley & Sons, U.S.A., distributed in the U.K. by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 96s.

"This handbook provides the electronic engineer with a reference of over 100 usable, well designed, reliable semiconductor circuits representing the most advanced techniques as well as design theory information.

Parts 2 through 10 were originally published by the Bureau of Ships, Department of the Navy, as 'A Handbook of Selected Semiconductor circuits' (NObsr 75231)."

The Mobile Manual for Radio Amateurs

Pp. 282. Published by The American Radio Relay League, West Hartford 7, Connecticut, U.S.A. Price $2.50 in the United States, $3.00 elsewhere.

"This revised edition of the Manual assembles under one cover

the most noteworthy articles on amateur mobile operation that have appeared in past issues of *QST*, the monthly journal of the League.

Included in the Manual for the first time are articles on emergency and portable equipment."

High Frequency Applications of Ferrites

By J. Roberts, B.Sc.(Eng.), A.C.G.I., A.M.I.E.E. Pp. 166 + x. Published by The English Universities Press Ltd., 102 Newgate Street, London, E.C.1. Price 16s.

"This book is intended primarily for electrical engineers and physicists who are engaged in the design or use of high frequency equipment.

The first chapter deals with those physical phenomena associated with ferrites which are essential for a clear appreciation of their properties. The rest of the book then surveys the application of ferrite components in communication and computer systems with particular reference to the principles of operation."

MEETINGS

I.E.E.

6*th March*. "An Investigation of the Usefulness of Back-Scatter Sounding in the Operation of H.F. Broadcasting Services", by E. D. R. Shearman, B.Sc.

14*th March*. "New Routes into National Certificate and Technician Courses", discussion to be opened at 6 o'clock by H. W. French, B.Sc.

21*st March*. "Self-Adaptive Control Systems", discussion to be opened by J. H. Westcott, B.Sc.(Eng.), Ph.D., J. L. Douce, M.Sc., Ph.D., P. E. W. Grinsted, Ph.D. and P. H. Hammond, B.Sc.

These meetings will commence at 5.30 with the exception of the one on 14th March, and will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2.

Brit. I.R.E.

16*th March*. "Transistorized Tape Recorders", by S. Welldon, at 5.30.

22*nd March*. "The Future of High Speed Storage Systems", discussion to commence at 5.30.

27*th March*. "Electronic Instrumentation for Cardiac Surgery", Symposium to be held at 3 o'clock at the Post Graduate Medical School, University of London, Hammersmith, W.12. Admittance by previous registration.

29*th March*. "Electronic Instrumentation for Nuclear Power Stations", Symposium to be held at 3 o'clock.

These meetings will be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, with the exception of the symposium on 27th March.

The Television Society

10*th March*. "Colour Television Camera Problems", by I. J. P. James, B.Sc.

23*rd March*. "Underwater Television in Marine Biology", by Dr. H. Barnes.

These meetings will be held at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2, and will commence at 7 o'clock.

Society of Relay Engineers

28*th March*. "Wired Distribution Networks for Television at Radio Frequencies", by L. E. Hooper-Smith, at 2.30 at the offices of B.I.C. Cables Ltd., 21 Bloomsbury Street, London, W.C.1.

ABSTRACTS INDEX

The Index to Abstracts and References included in the 1960 issues of *Electronic Technology* is now published.

Its cost is included in the annual subscription and it is thus despatched with this issue of *Electronic Technology* to all subscribers. It is available to non-subscribers at the price of 2s 6d, postage 4d.

"WIRELESS WORLD"

The April issue of this journal is to be published on 27th March, and with it *Wireless World* celebrates its 50th birthday. When it first appeared in April 1911, it bore the title *The Marconigraph* but changed to *Wireless World* in 1913.

With a life extending back to so early a time in the history of radio, there are few events and developments which have not been recorded in *Wireless World*. It is fitting, therefore, that it should mark its Golden Jubilee by including, among its April Contents, an article which traces the history of radio and electronic development from the near beginning.

ELECTROSTATIC PRINTING TUBES

Four electrostatic printing tubes that can translate electronic signals into printed words and pictures on paper have been introduced by Ratheon, U.S.A. The high-speed tubes can print 20,000 characters per second.

The tubes resemble flattened c.r.t.s with wire matrices across their faces. Tiny wires, 0·001 in. thick, spaced 250 to the inch, extend brush-like through the faces of the tubes. The varying cathode-ray beam current inside the tube passes through the wire matrix, depositing electrostatic charges on paper as it passes against the tube.

Pictures printed in this manner are said to have a resolution equalling that found in magazine illustrations.

The QV130 and QV131 tubes have 3-in. matrices, and have been designed for label or short copy printing.

The QV132 and QV133 are 10-in. printing tubes and can print on full-page sheets.

All tubes employ a medium persistence, green fluorescence and have magnetic deflection and focusing. Deflection angle is 40 degrees.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviations from nominal frequency for January 1961*

Date 1961 January	MSF 60 kc/s 1500 G.M.T. Parts in 10^9	Droitwich 200 kc/s 1030 G.M.T. Parts in 10^9
1	— 14·4	NM
2	— 13·7	+ 11
3	— 13·9	+ 11
4	— 13·8	+ 12
5	— 13·9	+ 12
6	— 14·1	+ 13
7	— 14·5	NM
8	— 14·4	NM
9	— 14·4	+ 16
10	— 14·3	+ 17
11	— 14·1	+ 18
12	— 14·8	+ 18
13	— 14·7	NM
14	— 14·6	NM
15	— 14·6	NM
16	— 14·6	+ 20
17	— 14·9	+ 20
18	— 14·6	+ 21
19	— 14·8	+ 20
20	— 14·8	+ 26
21	— 15·2	NM
22	— 14·9	NM
23	— 15·0	+ 22
24	— 14·9	+ 21
25	— 14·6	+ 25
26	— 14·9	— 22
27	— 14·9	— 20
28	— 15·2	NM
29	— 15·6	NM
30	— 15·4	— 21
31	— 15·2	— 16

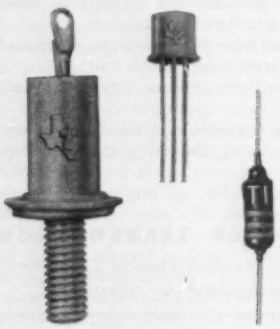
* *Nominal frequency corresponds to a value of 9 192 631 770 c/s for the caesium F,m(4,0) — F,m(3,0) transition at zero field. NM = Not Measured.*

New Products

Semiconductor Devices

Texas Instruments have recently announced eight new series of semiconductor devices, totalling over 80 products, which include gallium arsenide tunnel diodes for high-speed computer circuitry, silicon rectifier stacks for use in high-voltage power supplies, single-junction high-voltage silicon rectifiers, and germanium and silicon Mesa transistors.

Four gallium arsenide tunnel diodes (1N650 series) offer peak currents up to 10 mA ± 2%, large voltage swings, very high peak-to-valley ratios (greater than 15 : 1), forward voltages up to 1·1 V ± 5%, and operation up to 150 °C. They will operate at frequencies of at least 500 Mc/s.



The silicon-rectifier stacks 1N2878 series are available in 48 standard types with peak inverse voltages from 700 to 6,500 V. Twenty-four standard p.i.v. ratings are produced, each with either single-ended or axial-lead construction. At all 24 p.i.v. ratings, these units provide 250 mA average rectified forward current at 25 °C, 0·5 μA reverse current at p.i.v., and a low forward voltage drop.

The 1N1130 and 1N1131 devices are single-junction high-voltage silicon rectifiers with peak inverse voltage ratings of 1,500 V and average rectified forward current ratings of 300 mA at 25 °C and surge ratings up to 6·9 A for 10 msec. The device is housed in the 'high-hat' metal stud-mounting case, the 1N1130 having the cathode and the 1N1131 having the anode connected to the stud.

The 2G103, a diffused-base Mesa germanium transistor, will operate in saturated switching circuits up to 10 Mc/s and non-saturated circuits up to 50 Mc/s. It has a collector–base voltage rating of 15 V, current ratings of 50 mA at 25 °C case temperature, and a total device dissipation of 300 mW. Typical common-base alpha cut-off frequency is 300 Mc/s and h_{fe} (a.c. common-emitter forward current transfer ratio) is typically 6 dB at 100 Mc/s.

Double-diffused silicon Mesa transistors types 2N706A and 2N753 will operate in saturated switching circuits up to 10 Mc/s or up to 50 Mc/s in non-saturated current-mode switching circuits. Both types have a collector–base voltage rating of 25 V, total device dissipation at 25 °C case temperature of 1 W and a minimum h_{fe} (small-signal forward current transfer ratio) of 2, the 2N706A having an h_{FE} (d.c. beta) range of 20–60 and the 2N753, 400–120. The 2N753 is an improved high-gain version of the 2N706A.

Texas Instruments Ltd.,
Manton Lane, Bedford.

Miniature Capacitors

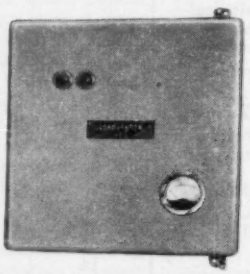
Johnson Matthey recently announced two Silver Star capacitors, types A15E and A25E, which are square in cross-section, of very small size and designed specifically for transistorized printed circuitry. The components are encapsulated in synthetic resin.

Type A15E is ¼ in. square × ½ in. long and has a capacitance range of 33–6,000 pF (minimum tolerance ± 2% or ± 2 pF below 100 pF). A25E has the same cross-section, but is 1 in. long and has a capacitance of 6,000–20,000 pF (minimum tolerance ± 1%). Both types are rated at 50-V peak working.

Johnson Matthey & Co. Ltd.,
*73–83 Hatton Garden, London, E.C.*1.

Continuous Level Indicator and Controller

Thomas Industrial Automation have developed a device known as the Levelrator which is designed to provide a continuous indication of the contents of silos, tanks, etc. and also operate visual and/or audible alarms when the contents reach pre-set high and low-level points.



The high and low-level control points can be set to operate over the full length of the sensing electrode and are completely independent of the indicating instrument which provides a continuous reading of the contents of the container.

The sensing electrode can be of the rigid variety for small vessels and the flexible type when the depth of the vessel exceeds 6 ft ; electrodes are available up to 60 ft in length.

Operating from standard mains supplies, the control unit consists of a stabilized power supply, alarm relay circuits and built-in calibrating instruments and it may be situated up to 150 ft from the electrode head unit.

Thomas Industrial Automation Ltd.,
Station Buildings,
Altrincham, Cheshire.

Microwave Power Generators

A Hewlett-Packard system of wide-band power generation at microwave frequencies has been announced by Livingston Laboratories. The frequency range 18–40 kMc/s is covered by two frequency doublers types 938A and 940A. Model 938A provides power from 18–26·5 kMc/s when driven by a 9–13·25-kMc/s signal. Model 940A provides power from 26·5 to 40 kMc/s when driven by a 13·25–20-kMc/s signal.

The frequency-doubler units consist essentially of crystal harmonic generators suitably mounted in a waveguide. Also incorporated are a power monitor and a precision attenuator. Both units may be driven by klystrons, signal generators or sweep oscillators. Since both models are broadband instruments, the input signal may be c.w., pulsed or swept. The output signal retains the facilities of the driving source.

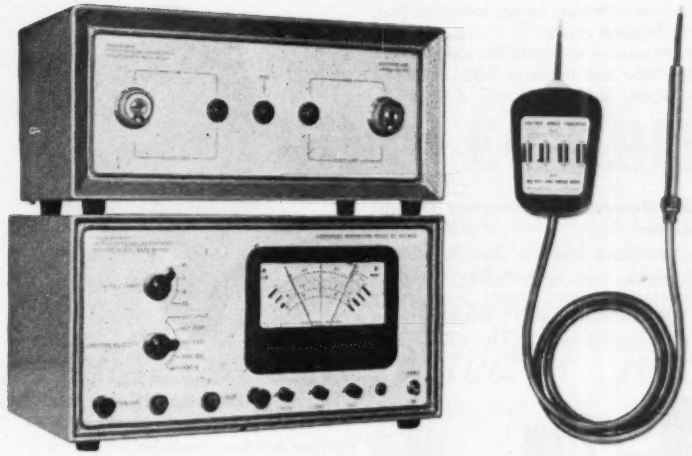
Both frequency doublers have built-in power monitors and precision rotary-vane attenuators which allow accurate level setting from 0 to −100 dB with an absolute level accuracy of ± 1 dB at most frequencies. The incremental attenuator accuracy is 2% of the reading or ± 0·2 dB, whichever is the greater, and the output v.s.w.r. is of the order of 2 : 1 at maximum output and less than 1·2 : 1 with the attenuator set at −10 dB or more.

Distributed in the U.K. by

Livingston Laboratories Ltd.,
*Retcar Street, London, N.*19.

High-Speed Component Comparator

A high-speed component comparator, designed primarily for testing resistors, capacitors and inductors under mass-production conditions, has been produced by British Physical Laboratories. Known as the Component Comparator Model CZ457 Mk. II, it directly compares the component under test with an external standard unit and indicates the percentage deviation of the value of the unknown component within the



pre-set tolerance limits ±1·5%, ±12% or ±22%.

Using one external standard *R*, *C* or *L* for the measurement of percentage deviation, the component ranges are 10 Ω to 15 MΩ, 20 pF to 10 μF, and 2 mH to 100 H for resistors, capacitors, and inductors, respectively.

When using the instrument for direct measurement of *R*, *C* and *L*, two external standards are used and the ranges covered are 0·01 Ω to 2·2 MΩ, 0·02 pF to 1·8 μF, and 2 μH to 18 H.

The accuracy of the unit is ±3% of indication deviation or 1% f.s.d., whichever is the greater, and the maximum theoretical test speed is 6,000 components per hour.

B.P.L. (Instruments) Ltd.,
Radlett, Herts.

High-Sensitivity 3½-in. C.R.T.

The M.O. Valve Co. have introduced a 3½-in. diameter high-sensitivity cathode-ray tube with helical p.d.a. It is designed for use in compact oscilloscopes which need a flat-faced tube.

Typical deflection sensitivities are 5·7 V/cm and 30 V/cm for the X and Y plates, respectively.

The tube is available with various screen phosphors.

The M.O. Valve Co. Ltd.,
Magnet House,
Kingsway, London, W.C.2.

Voltage-Range Neon Indicator

Martindale Electric have developed a voltage-range indicator which uses four neon tubes and can be used in the 50–600-V a.c. or d.c. range. In operation the four neons light progressively as the voltage rises, neon 1 operating at about 50 V a.c. and 60 V d.c.; neon 2 at about 100 V a.c. and 140 V d.c.; neon 3 at about 200 V a.c. and 240 V d.c., and neon 4 at about 400 V a.c. and 560 V d.c. It is stated that the instrument will operate at currents below the 'threshold of pain' level (somewhat around 5 mA at 500 V) but is not so sensitive that it will record stray currents produced by intercapacitance and other causes.

The instrument has one retractable shrouded prod projecting from the case and a wander lead. At the end of the lead is a probe with a retractable insulated sheath covering a metal prod. The probe holder contains a fuse and limiting resistor.

Measuring approximately 3½ in. × 2½ in., the instrument is housed in a robust plastic case.

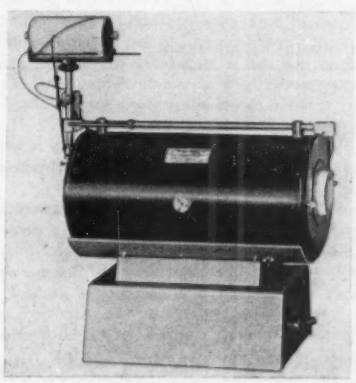
Martindale Electric Co. Ltd.,
Westmorland Road, London, N.W.9.

Laboratory Furnaces

A range of precision laboratory furnaces with a temperature stability of ±0·25 °C, an interchangeable drum control for thermal cycling, and facilities for heat-loss compensation at the ends of the tube have recently been introduced by Shandon Scientific.

The standard range comprises three models for maximum temperatures of 1050 °C with tube size respectively of 35 mm i.d. × 290 mm length (for operation on 115 V a.c.), 60 mm × 450 mm (250 V a.c.) and 75 mm × 700 mm (220 V a.c.); and two models for maximum temperatures of 1250 °C, with tube sizes of 35 mm × 290 mm (115 V a.c.) and 60 mm × 450 mm (220 V a.c.). Two special models for maximum temperatures of 1500 °C are also available, with tube sizes of 16 mm × 250 mm (115 V a.c.) and 33 mm × 400 mm (220 V a.c.).

Each of the standard models may be controlled either by a thermostatic regulator or a chronograph drum control for thermal cycling.



The thermostatic regulator consists of a graduated scale and a pointer operated by the expansion of a wire running through the furnace body. The scale is fitted with a stop which may be set to any desired temperature. When the pointer reaches the stop, the current is automatically cut off from the heating windings via the 8-V relay.

The chronograph drum control for thermal cycling is clockwork-operated. The drum carries a temperature chart on which the cycling programme is marked out by a strip of thick cardboard. In operation, the drum revolves at one of two set speeds and, as the furnace heats up, a contact on an arm connected to the expansile wire moves across the drum. When it meets the cardboard, the relay operates and the current is cut off. In this way the contact follows closely the edge of the cardboard strip, alternately making and breaking the circuit to the windings, and thus holding the temperature to the planned programme.

Three separate windings are mounted side by side along the tube and may be controlled individually, thus enabling the temperature to be varied along the tube.

Shandon Scientific Co. Ltd.,
6 Cromwell Place, London, S.W.7.

Proximity Industrial Detector Unit

A transistorized proximity detector which signals the presence or absence of a metallic object in a given region of space is the latest development of Mec-Test Ltd.

The equipment, known as Proximitron Type S, consists of a sensing element and a control unit. A variety of sensing elements is available including loop coils (which may be buried in the ground), annular coils and probes, which will detect the presence or position of both ferrous and non-ferrous metals. The field of detection can range from a fraction of an inch to several feet according to the type of probe selected.

In operation, one coil in the sensing element is energized with a 2-kc/s signal and a second coil, also in the sensing element, is used to detect any change in the magnetic field brought about by the presence of metal in the field. The signal from the detecting coil is amplified and

detected and may be used to close a relay or to provide a d.c. output pulse.

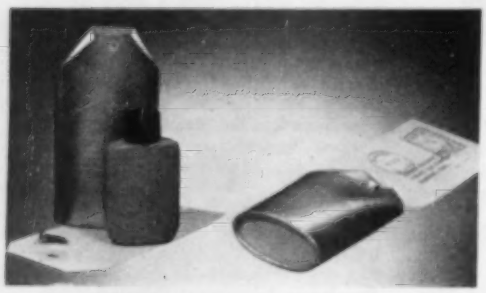
Mec-Test Ltd.,
Henrietta House,
Henrietta Place, London, W.1.

Component Postal Pack

Rolex Paper Co. have recently introduced a lightweight pack for mailing small samples.

It is known as the Rol-Sac and, in fact, resembles a small sack but is made of laminated cardboard with the neck being held closed by a brass-head type of paper clip. The pack can be built with a range of barrier liners including polythene and aluminium foil which can be wound on the inside or buried in the wall as an inner ply.

Rolex Paper Co. Ltd.,
21-24 Chiswell Street, London, E.C.1.



Double-Beam Oscilloscope

An improved version of the Solartron oscilloscope type CD1014 with additional trigger modes and provision for external brightness modulation has recently been announced; it is known as portable double-beam oscilloscope type CD1014.2.

Two identical Y amplifiers with a bandwidth d.c. to 5 Mc/s and a maximum sensitivity of 100 mV/cm are used. A built-in pre-amplifier preceding Y_2 provides an increased sensitivity of 1 mV/cm at reduced bandwidth. The inputs can be 'floating' with respect to chassis.

The basic time-base range, 1 cm/sec to 1 cm/μsec ($\pm$5% measuring accuracy) may be extended to 5 cm/μsec with the X expansion and the oscilloscope may be triggered internally or externally (positive and negative) on television frame, television line, Auto or Normal modes.

The instrument weighs 22 lb., incorporates a 3½-in. tube and operates from a normal mains input of 90–130 and 200–240 V, 45–66 c/s.

Solartron Laboratory Instruments Ltd.,
Cox Lane, Chessington, Surrey.

Medium-Power High-Linearity Output Transistor

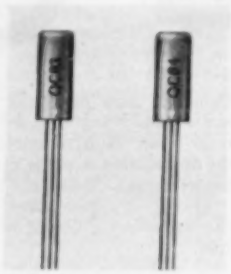
Mullard have recently introduced the OC83 which is a medium-power output transistor of the germanium p-n-p alloy type for industrial a.c. amplifier and switching applications. (The OC83 is a lower-cost alternative to the recently-introduced OC84.)

The device has a peak current of 1 A and a high linearity of current gain; the minimum gain being 50 at 50 mA and 40 at 300 mA.

Absolute maximum voltage ratings for the OC83 include a collector-to-base voltage of -32 V, and a collector-to-emitter voltage of -32 V. In addition, the transistor is designed to withstand a collector-to-emitter voltage of -20 V for all currents up to 300 mA.

The OC83 has a junction temperature rating of 85 °C which, coupled with the thermal resistance of 0·25 mW/°C in free air, makes possible a dissipation of 160 mW in an ambient temperature of 45 °C. This can be increased to 400 mW with a suitable heat sink. The typical cut-off frequency (f_1) of the device is 850 kc/s.

Mullard Ltd.,
Torrington Place, London, W.C.1.



Angular Displacement Micrometer

A unit which has been designed to provide a quick and easy method of setting an angular displacement, to the nearest minute of an arc, on a variety of small electrical components such as synchros and potentiometers, has been announced by the Autotest Division of Elliott Bros.

Known simply as Micrometer Dial it has an accuracy of the order of $\pm$ 15 seconds of arc.

The instrument consists of a light alloy body containing the main drive and a turns-counting mechanism, a dial marked from 0 to 360°, a micrometer drive marked off in minutes of arc, and an adaptor plate.

The adaptor plate can be supplied to mount either a 10-turn helical potentiometer or any size synchro.

The main drive gear is designed in such a way that wear on the teeth is automatically compensated, thus maintaining its accuracy.

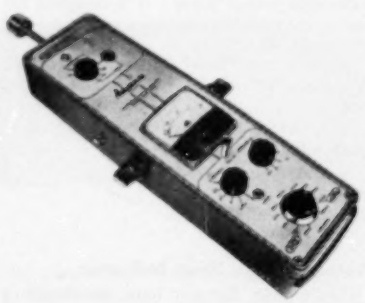
Elliott Brothers (London) Ltd.,
Autotest Division,
Airport Works,
Rochester, Kent.

Sound Spectrometer

Advance Components have negotiated a licence with Peekel Laboratorium voor Electronica N.V. of Holland to manufacture under licence their sound spectrometer Type GRB in the U.K. The instrument will be known as the Advance Sound Spectrometer Type SPM.1.

Conforming fully to the latest international requirements of the Rapallo Convention, this instrument is compact (17 in. × 4·75 in. × 3·2 in.), of comparatively light weight (9 lb) and, with battery power supply and microphone incorporated, the unit is self-contained and truly portable.

Noise measurements can be made in phons or decibels with an accuracy to A.S.A. and D.I.N. standards. The overall frequency coverage is from 20 c/s to 12 kc/s and sounds within this range can be analysed in separate frequency bands by suitable selection from one of eight filters.



Provision is made for the connection of a transducer or accelerometer for vibration analysis; measurements, including internal battery checks, are made on a built-in moving-coil meter. A jack socket permits the use of an external monitoring device.

The total measuring range of the instrument is from 20 dB to 150 dB with respect to a sound-pressure reference of 2×10^{-4} dynes/cm^2.

Advance Components Ltd.,
Roebuck Road,
Hainault, Ilford, Essex.

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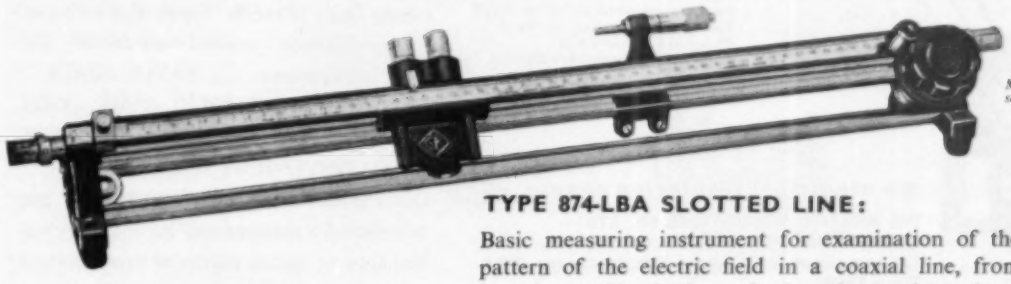
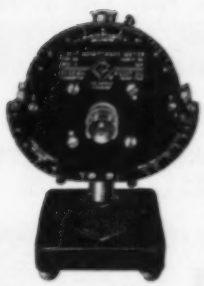
Complete · Integrated · Accurate · Versatile



Micrometer Vernier shown attached.

TYPE 874-LBA SLOTTED LINE:

Basic measuring instrument for examination of the standing-wave-pattern of the electric field in a coaxial line, from which VSWR, impedance of load, phase of reflected wave, losses in attached elements, degree of mis-match between line and load, etc. can be determined. Accurate and straightforward in use. Frequency range 300-5000 Mc/s (with some loss in accuracy: 150-7000 Mc/s). Also available: Type 874-LV Micrometer Vernier (for measurement of high VSWR). Type 874-MD Motor Drive for oscillographic display of standing-wave pattern.



TYPE 1602-B ADMITTANCE METER:

A compact and versatile instrument, accurate and rapid in use, for determining the components of an unknown admittance in the VHF-UHF range. Scales read directly in conductance and susceptance, *independent of frequency*. With unknown connected through quarter-wavelength line, scales read in resistance and reactance. Can be used for measurement of VSWR and reflection coefficient, matching or comparison of impedances, and measurements on balanced line circuits (with the Type 874-UB 'Balun'). Frequency range 41-1500 Mc/s. (Down to 10 Mc/s, with correction). Includes conductance and susceptance standards.

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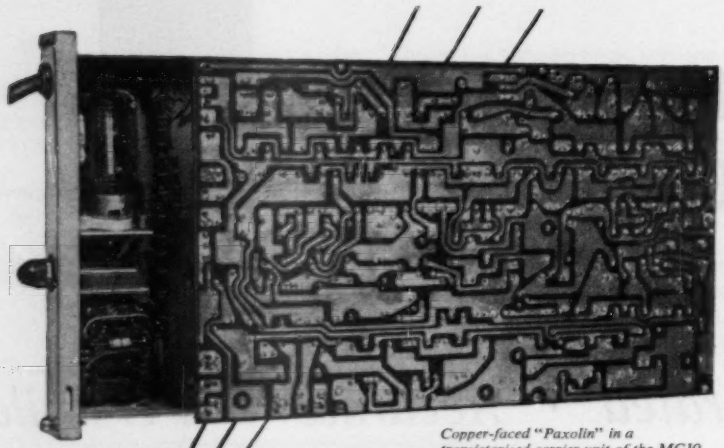

Copper-faced "Paxolin" in a transistorised carrier unit of the MC10 V.H.F. Radio Telephone Terminal. Photograph by courtesy of A.E.I. Telecommunications Division.

Lines of Communication

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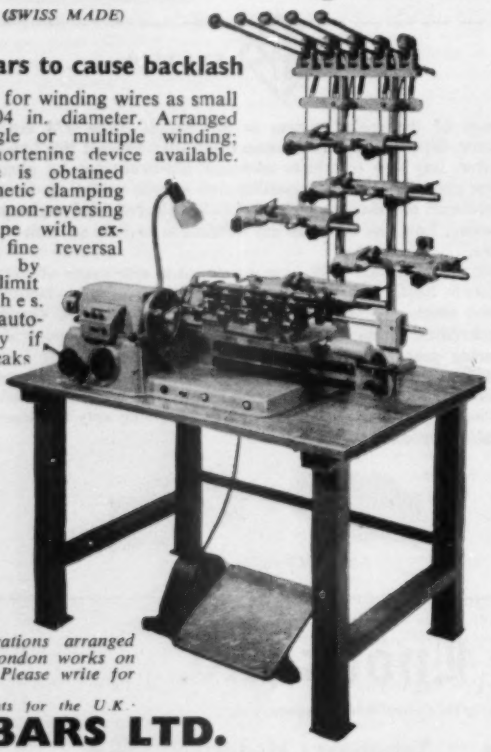

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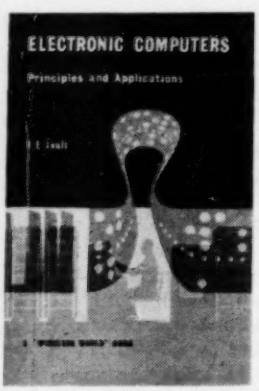

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Principles and Applications

by T. E. Ivall

SECOND EDITION

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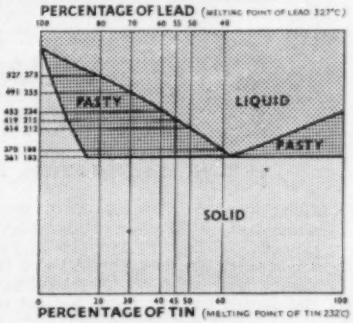
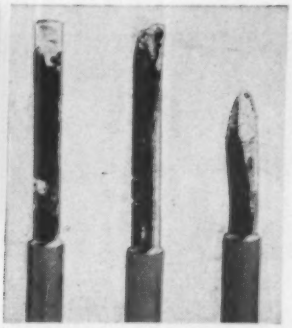
CONSTITUTION OF ALLOYS OF ERSIN MULTICORE SOLDER

The diagram shows that all the standard alloys of Ersin Multicore Solder have a plastic range, i.e., on heating they are pasty between the solid and liquid states. Practical experience has shown that there are advantages in having a plastic range, e.g. for tag jointing where the use of 60/40 alloy obviates fractures which may occur with other alloys where there is slight vibration while the solder is setting solid.





(left) after 10,000 joints with Savbit. (centre & right) after 1,000 and 7,500 joints respectively with standard tin/lead alloy.

PRINTED CIRCUITS

Leaflet P.C.L. 101 gives full details of a complete soldering process developed by Multicore Laboratories for printed circuits.

SAVBIT TYPE 1 ALLOY

Made under sole British Licence of Patent No. 721,881.

Savbit Type 1 alloy was developed after extensive research in the Multicore Laboratories into the main causes of bit wear. It incorporates a small percentage of copper which prevents absorption of copper from the bit into the solder alloy. After prolonged tests, it was found that the life of solder bits was increased by up to 10 times. The speed of soldering is not affected.

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L.M.P. alloy, with 2% silver content, which melts at 179°C for silver-coated components.

PUBLICATIONS

Laboratory engineers and technicians are invited to write on their company's letter-heading for the latest edition of Modern Solders. It contains data on melting points, gauges, alloys, etc.

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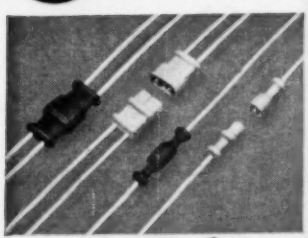
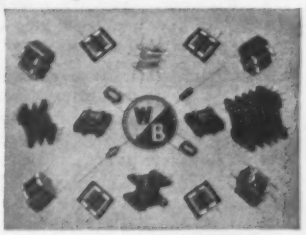


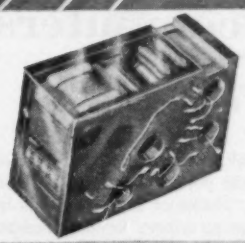
PLUGS AND SOCKETS

The plugs and sockets shown are for use in underwater junctions or outside installations. Moulded in Alkathene the loading capacity is 350 volts at 5 amps. max.

MINIATURE BOBBINS

Illustrated are a range of moulded bobbin assemblies for small transformers and chokes. Moulded in Polystyrene or Polythene, they are available in various colours.





P.O. TRANSISTOR AMPLIFIER

This P.O. Transistor Amplifier moulding is another example of potted components, the illustration is of a development of a miniature version of the amplifier 100A, this model utilising transistors to replace the miniature valves.

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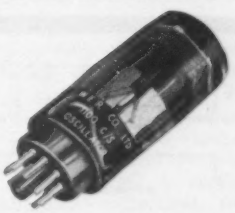

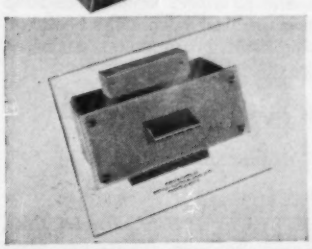

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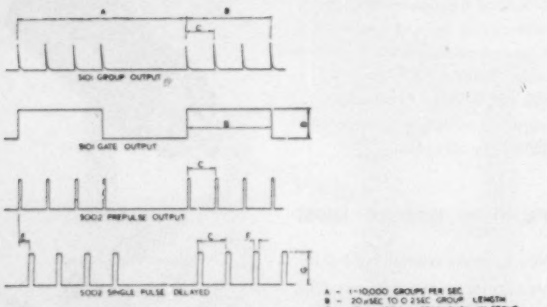

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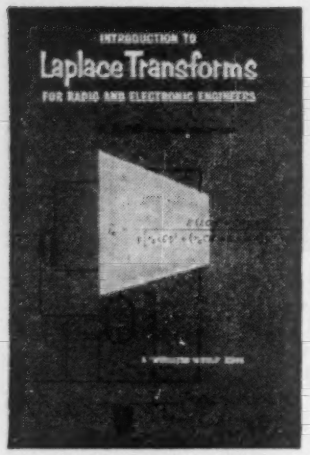

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